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The measurement of system-impulse response by means of cross-correlation with binary signals

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Summary

The measurement of system-impulse responses by means of cross-correlation is discussed. With respect to the instrumentation of those measurements classes of binary signals offer appreciable advantages when used as test signal.

The smoothing time for both periodic and non-periodic binary signals is determined. From this it is found that the use of periodic-binary signals leads to an appreciable reduction of the smoothing time (i.e. a smaller variance of the resulting information after a certain interval of correlation).

In addition to that advantage such a type of test signal, — together with the same signal delayed over an arbitrary interval —, can be generated in a simple way by a shift register with adders and feedback. Finally the effects of additive noise on the smoothing time is studied.

1. Correlator theory

When a linear process is excited with a stochastic input, there will be correlation between the system output and the input. The amount of correlation depends upon the nature of the system and of the system input.

Consider now a linear system, having an impulse response $h(t)$ and a stationary random input $x(t)$.

The system output is easily obtained from the convolution integral as [1])

$$y(t) = \int_{0}^{\infty} x(t-\lambda)\, h(\lambda)\, d\lambda \qquad (1)$$

The crosscorrelation function of $x(t)$ and $y(t)$ is defined as the expectation of their product. Thus

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$$\Psi_{xy}(\tau) = E[x(t)\, y(t-\tau)] \tag{2}$$

For stationary inputs the order in which time and ensemble averages are taken can be interchanged and this leads to the result

$$\psi_{xy}(\tau) = \int_0^\infty \psi_{xx}(\tau - \lambda)\, h(\lambda)\, d\lambda \tag{3}$$

where $\psi_{xx}(\tau)$ is the autocorrelation function of the input.

Consider we have an input signal $x(t)$ with an autocorrelation funtion

$$\psi_{xx}(\tau) = N\,\delta(\tau)$$

where $\delta(\tau)$ is the dirac-funtion.

Substitution in (3) leads to:

$$\psi_{xy}(\tau) = \int_0^\infty N\,\delta(\tau - \lambda)\, h(\lambda)\, d\lambda = Nh(\tau) \tag{4}$$

Hence it is clear that the determination of the system-impulse response can be achieved by measuring the crosscorrelation between the system output and a random input with suitable autocorrelation function.

So the problem is to find an input signal with an autocorrelation function which approximates the delta function as closely as possible.

2. The use of stationary binary signals

The advantages of binary signals

In almost any case we may use as correlator input any stationary random signal with great bandwidth.

In such cases a considerable practical advantage can be gained by allowing the test signal $x(t)$, to have only two values, say $\pm a$. In the first place the delay of such binary signals is considerably easier than that of continuous signals and it can be achieved with one of the available digital storage devices.

The multiplication operation is much easier to mechanize because the binary signal reduces it to a simple gating operation. Another advantage of a binary signal is that it contains the maximum energy for a given peak value. Thus it is possible to obtain the greatest output signal-to-noise ratio for a given degree of system disturbance.

This may be important if the crosscorrelation must be achieved without disturbing normal system operation or without driving it into saturation.

2.1. *Types of binary signals*

Although there are many types of binary signals, only two types will be discussed here.

a. The random telegraph signal.
   The sign changes occur at random times and are independent. The number of sign changes in a long time interval T is assumed to obey a Poisson distribution.

b. The discrete interval binary signal.
   The sign changes occur now at discrete time instants, which are separated by some integral multiple of the minimum interval t_1.

The state in each interval is independent of the state in any other interval.

2.2. *Crosscorrelation with the random telegraph signal*

The autocorrelation function of this type of signal can be shown to be [2]).

$$\psi_{xx}(\tau) = a^2 \varepsilon^{-2\mu|\tau|} \tag{5}$$

where μ is the average number of sign changes per second and a is the peak value of the signal.

Substitution in (3) leads to (fig. 1):

$$\psi_{xy}(\tau) = a^2 \int_0^\infty \varepsilon^{-2\mu|\tau-\lambda|} h(\lambda)\, d\lambda \tag{6}$$

The impulse response $h(t)$ of a linear system with single poles can be given in general as:

$$h(\tau) = \sum_{m=0}^{M} A_m \varepsilon^{-a_m t} + \sum_{n=0}^{N} B_n \varepsilon^{-a_n t} \sin(\omega_n t + \varphi_n) \tag{7}$$

where A_m and B_n are constant coefficients.
Substitution in (6) leads at last to:

$$\psi_{xy}(\tau) = a^2 \sum_{m=0}^{M} A_m \left(\frac{\varepsilon^{-a_m \tau}}{\mu} - \frac{\varepsilon^{-2\mu\tau}}{2\mu} \right) +$$

$$+ a^2 \sum_{n=0}^{N} B_n \left\{ \frac{4u}{\omega_n^2 + 4\mu^2} \varepsilon^{-a_n \tau} \sin(\omega_n \tau + \varphi_n) + \right.$$

$$\left. + \left(\frac{\omega_n \cos\varphi - 2\mu \sin\varphi}{\omega^2 + 4\mu^2} \right) \varepsilon^{-2\mu\tau} \right\} \quad . \ . \ . \ \text{if } \tau \leqq 0$$

$$\psi_{xy}(\tau) = a^2 \sum_{m=0}^{M} A_m \frac{\varepsilon^{2\mu\tau}}{2\mu} + a^2 \sum_{=0}^{N} B_n \frac{2\mu \sin\varphi_n + \omega_n \cos\varphi_n}{\omega_n^2 + 4\mu^2} \varepsilon^{2\mu\tau}$$

$$\text{if } \tau \geqq 0$$

with the additional condition $a_n << 2\mu$.

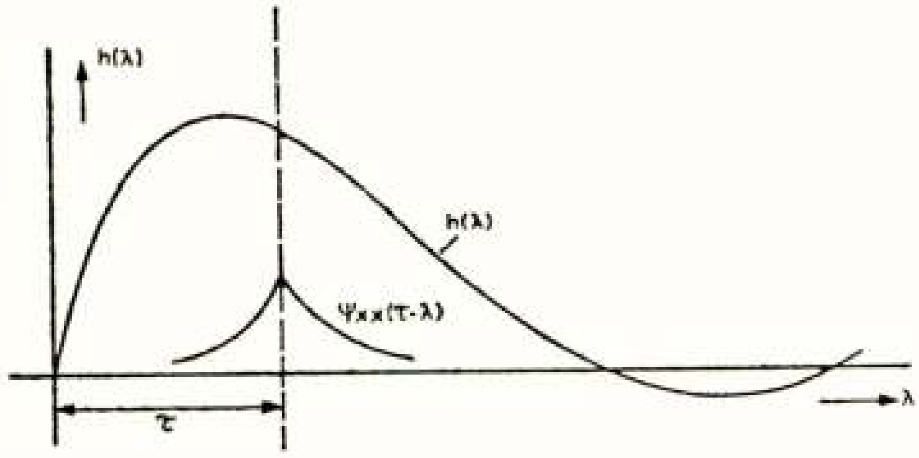

Fig. 1
Graphical representation of the convolution

Comparing these results with equation (7) we see that there exists:

1. an error depending on τ, so it is impossible to determine $h(t)$ when $\mu\tau << 1$.
2. a systematic error which can be made small if we choose $\omega_n << 2\mu$. This means that the input noise spectrum must be much wider than the system bandwidth.

2.3. *Crosscorrelation with discrete interval binary noise*

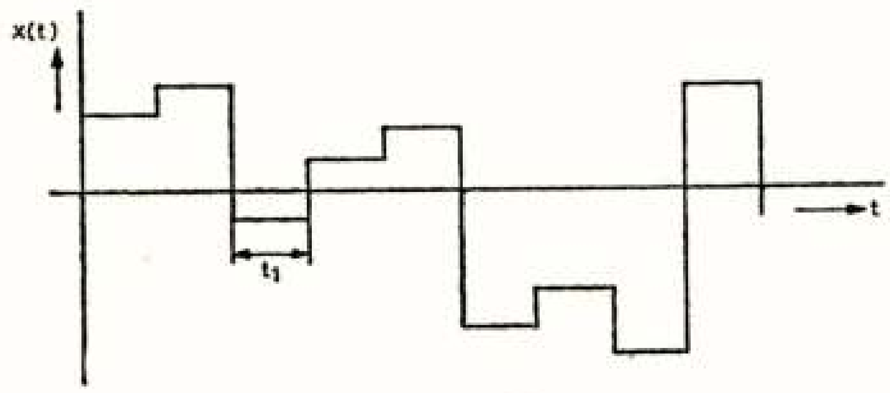

Fig. 2
A discrete interval noise

As a second example, the wave shown in fig. 2 is considered. The signal is constant for a time interval of t_1 seconds and then jumps to the other value. The probability that the signal is positive is equal to the probability that the signal is negative.

If A_n is the magnitude of the signal $f(t)$ in the interval $\{n t_1, (n+1) t_1\}$ the probability that A_n lies between x and $x + dx$ is given by

$$P(x < A_n < x + dx) = p(x)\, dx \tag{8}$$

The autocorrelation function $\psi(\tau)$ can be computed from:

$$\psi(\tau) = \iint e_1 e_2 p(e_1, e_2, \tau)\, de_1\, de_2$$

Here $p(e_1, e_2, \tau)$ is the probability of $f_1(t)$ lying between e_1 and $e_1 + de_1$ at a given time and τ seconds later between e_2 and $e_2 + de_2$.

The simplest procedure involves breaking the integrand into two parts, one representing the contribution when $e_1 = e_2$ (or t and $t + \tau$ lie in the same interval) and the other when t and $t + \tau$ lie in different intervals.

This leads to:

$$\psi(\tau) = \iint_{-\infty}^{+\infty} x.\, x.\, p(x, x, \tau)\, dx\, dx + \iint_{-\infty}^{+\infty} x\,.\, y\, p(x, y, \tau)\, dx\, dy \qquad (9)$$

The probability that e_1 lies between x and $x + dx$ is $p(x)\,dx$ and the probability that $e_1 = e_2$ is simply the probability that t_1 and $t_1 + \tau$ lie in the same interval, which is evidently $1 - \frac{|\tau|}{t_1}$ if $\tau < t_1$ and zero if $\tau > t_1$, so

$$p(x, x, \tau) = \left(1 - \frac{|\tau|}{t_1}\right) p(x) \qquad (10)$$

The limits of integration are changed from $-\infty$ and $+\infty$ to 0 and $+\infty$ because equation (8) refers to the magnitude without question of sign.

The second term in (9) vanishes identically since the values of A_n in any two different intervals are independent and each has an average value of zero.

Hence

$$\left.\begin{aligned} \psi(\tau) &= \left(1 - \frac{|\tau|}{t_1}\right) \int_0^\infty x^2 p(x)\, dx \qquad && \text{if } |\tau| \leq t_1 \\ \psi(\tau) &= 0 && \text{if } |\tau| > t_1 \end{aligned}\right\} \qquad (11)$$

So for each probability distribution function $p(x)$ for which holds:

$$\int_0^\infty x^2 p(x)\, dx = b^2 < \infty$$

the autocorrelation function can be written as:

$$\left.\begin{aligned} \psi(\tau) &= b^2\left(1-\frac{|\tau|}{t_1}\right) && \text{if } |\tau| \leqq t_1 \\ \psi(\tau) &= 0 && \text{if } |\tau| > t_1 \end{aligned}\right\} \tag{12}$$

If we have a signal of which the amplitude has only two values, say $+a$ and $-a$ then:

$$p(x) = \delta(x-a)$$

The autocorrelation function is then:

$$\left.\begin{aligned} \psi_{xx}(\tau) &= a^2\left(1-\frac{|\tau|}{t_1}\right) && \text{if } |\tau| \leqq t_1 \\ \psi_{xx}(\tau) &= 0 && \text{if } |\tau| > t_1 \end{aligned}\right\} \tag{13}$$

Substitution of this result in equation (3) leads to (fig. 3):

$$\psi_{xy}(\tau) = a^2 \int_0^\infty \left(1 - \frac{|\tau-\lambda|}{t_1}\right) h(\lambda)\, d\lambda \tag{14}$$

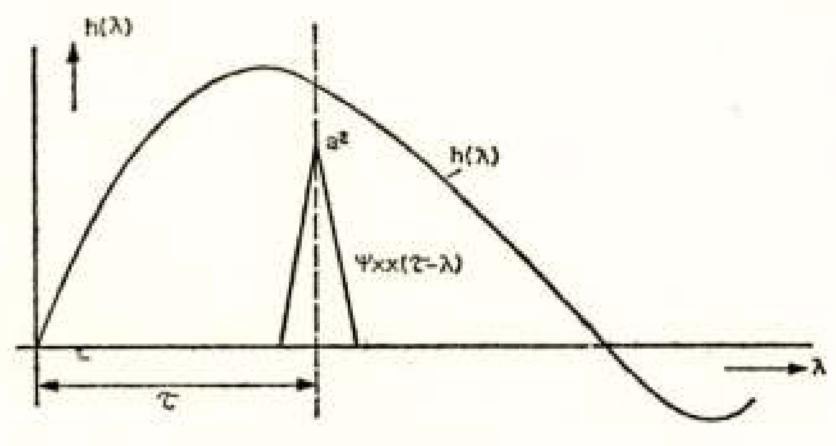

Fig. 3
Graphical representation of the convolution

The calculation of the error may be done in the same way as in section 2.4. To minimize the error the correlator input signal must satisfy the following conditions:

a. the duration t_1 of the smallest pulse must be much smaller than the time constant of the process.
b. the time t_1 must be much smaller than the oscillation time of the process.

Another advantage of the discrete interval binary noise is that the error for $\tau > t_1$ does not depend on the time delay τ and that the delay may be obtained easily with a shift register.

As a final result for the impulse response we obtain:

$$h(\tau) = \frac{\psi_{xy}(\tau)}{t_1 \psi_{xx}(0)} = \frac{\psi_{xy}(\tau)}{a^2 t_1}$$

The basic correlator is given in fig. 4.

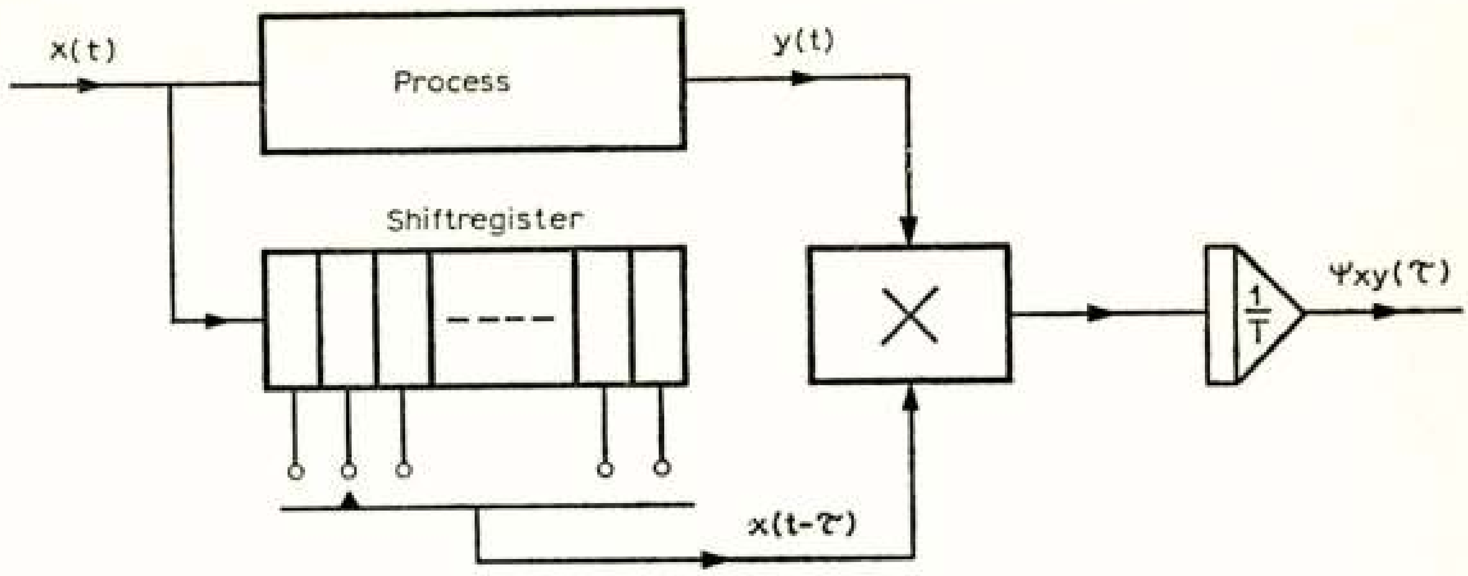

Fig. 4
The basic correlator

3. The use of periodic binary signals

Under certain circumstances [3][4]) it is possible to generate periodic binary signals, called maximum length sequences, using a shift register with "modulo 2" feedback.

The autocorrelation function of these signals has the character of the autocorrelation function of discrete interval binary noise. However, the autocorrelation function of the first signal is periodic, with a period in the time delay variable equal to the period of the signal itself.

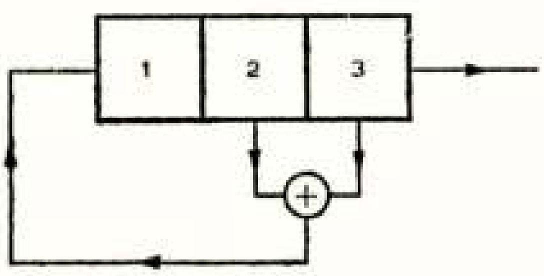

Fig. 5
Shift-register with modulo 2 feedback

As an example we choose a shift-register with three sections as shown in fig. 5. Starting the register e.g. with a one in section 1 and a zero in sections 2 and 3, then the output of the register is as shown in fig. 6.

We see that after seven steps the register is back at the starting point.

One period of the generated sequence may be written as 1 0 1 1 1 0 0.

The corresponding pulse series is shown in fig. 7.

The output of section 2 and 3 is equal to the output of section 1 but delayed over a time $\frac{T}{7}$ viz. $\frac{2T}{7}$.

The autocorrelation function of a binary sequence is proportional to the number of ones that correspond when two periods

1	2	3
1	0	0
0	1	0
1	0	1
1	1	0
1	1	1
0	1	1
0	0	1

Fig. 6
Binary representation of the section outputs of the shift-register shown in fig. 5

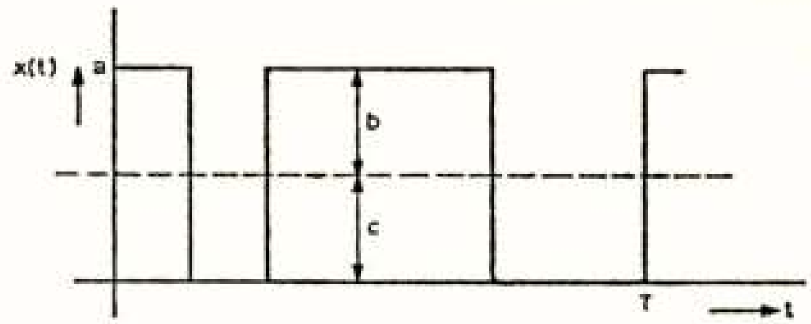

Fig. 7
The generated sequence of the generator shown in fig. 5

of the generated sequence, one delayed with respect to the other, are placed one below the other and compared in vertical sense.

The autocorrelation function at the time delay $\tau = \frac{nT}{7}$ is now equal to:

$$\begin{aligned} \psi\left(\frac{nT}{7}\right) &= \frac{4}{7}\, a^2 \qquad \text{if} \qquad n = 0 \text{ or } 7 \\ &= \frac{2}{7}\, a^2 \qquad \text{if} \qquad 1 \leqq n \leqq 6 \end{aligned} \qquad (15)$$

It is easily proved that the autocorrelation function $\psi(\tau)$ in the interval $\frac{T}{7} < \tau < \frac{6T}{7}$ is independent of τ and equal to $\frac{2}{7}\, a^2$.

Fig. 8
Autocorrelation function of the sequence shown in figure 7

In the intervals

$$\left(pT - \frac{T}{7}\right) \leqq \tau \leqq \left(pT + \frac{T}{7}\right)$$

where p is an arbitrary integer (positive or negative), $\psi(\tau)$ is changing linearly with τ (fig. 8). In general it can be shown[5]) that a shift-register with n sections with "modulo 2" feedback, such that a max. length sequence with period time T is generated, the autocorrelation function may be written as:

$$\psi(\tau) = \frac{2^{n-2}}{2^n - 1}\, a^2 \left(2 - \frac{|\tau|}{t_1}\right) \qquad \text{if} \qquad pT - t_1 \leqq \tau \leqq pT + t_1$$

where $t_1 = \dfrac{T}{2^n - 1}$ and $p = \ldots -3, -2, -1, 0, 1, 2, 3, \ldots$

$$\text{and } \psi(\tau) = \frac{1}{2}\,\psi(0) \text{ elsewhere.}$$

To approximate the autocorrelation function of discrete interval binary noise it is necessary that $\psi(\tau)$ equals zero outside the intervals $pT - t_1 \leqq \tau \leqq pT + t_1$.

To obtain this we choose the values of b and c in fig. 7 equal to:

$$\left.\begin{aligned} c &= \frac{1}{2^n - 1}\left\{2^{n-1} \pm \sqrt{2^{n-2}}\right\} a \\ b &= \frac{1}{2^n - 1}\left\{2^{n-1} - 1 \mp \sqrt{2^{n-2}}\right\} a \end{aligned}\right\} \tag{16}$$

The general expression of the autocorrelation function $\psi(\tau)$ of a maximum length sequence under this condition is at last:

$$\psi_{xx}(\tau) = \frac{2^{n-2}}{2^n - 1}\, a^2 \left\{1 - \frac{|\tau|}{t_1}\right\} \qquad \text{if}$$

$$(pT - t_1) \leqq \tau \leqq (pT + t_1) \qquad \text{and}$$

$$\psi_{xx}(\tau) = 0 \text{ elsewhere} \tag{17}$$

The solution of the convolution integral (3) may be obtained in the same way as in section 2.2. There is however an additional restriction viz. that the impulse response has to be damped away within the time of one period of the sequence (fig. 9).

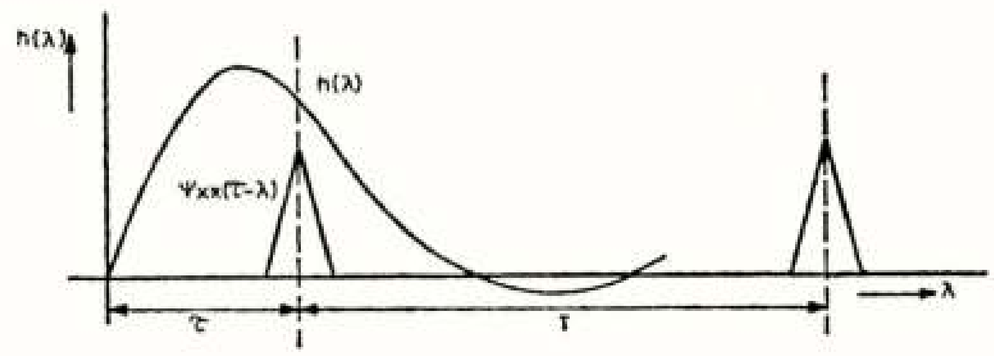

Fig. 9
Graphical representation of the convolution

As a final result for the impulse response we obtain:

$$h(\tau) = \left\{\frac{2^n - 1}{2^{n-2}}\right\}^2 \frac{\psi_{xy}(\tau)}{a^2\, T} \tag{18}$$

A second important property of the maximum length sequence is that when we add "modulo 2" sequences that are delayed with respect to each other we obtain a third sequence delayed with respect to the two other sequences [3]).

By proper combination and modulo 2 addition of the available section outputs, it is possible to obtain all the possible phases of the maximum length sequence in one period. So it is very simple to mechanize the time delay of the generated signal.

As an example we choose again a shift-register with three sections. The possible sequences and the positions of the switches *a*, *b* and *c* to obtain them are given in fig. 10.

Phase	Sequence							Position of switches a	b	c
0	1	0	1	1	1	0	0	1	0	0
1	0	1	0	1	1	1	0	0	1	0
2	0	0	1	0	1	1	1	0	0	1
3	1	0	0	1	0	1	1	1	0	1
4	1	1	1	0	1	0	1	1	1	1
5	1	1	1	0	0	1	0	1	1	0
6	0	1	1	1	0	0	1	0	1	1

Fig. 10
Possible output sequences with corresponding switch positions

If we consider now the successive positions of one switch we notice another maximum length sequence, equal to the generated sequence, but in reversed direction. This property is generally valid. The block diagram of the correlator as finally mechanized is shown in fig. 11.

4. The influence of noise on the measurements

4.1. *Introduction*

In the previous discussions we have assumed that the system

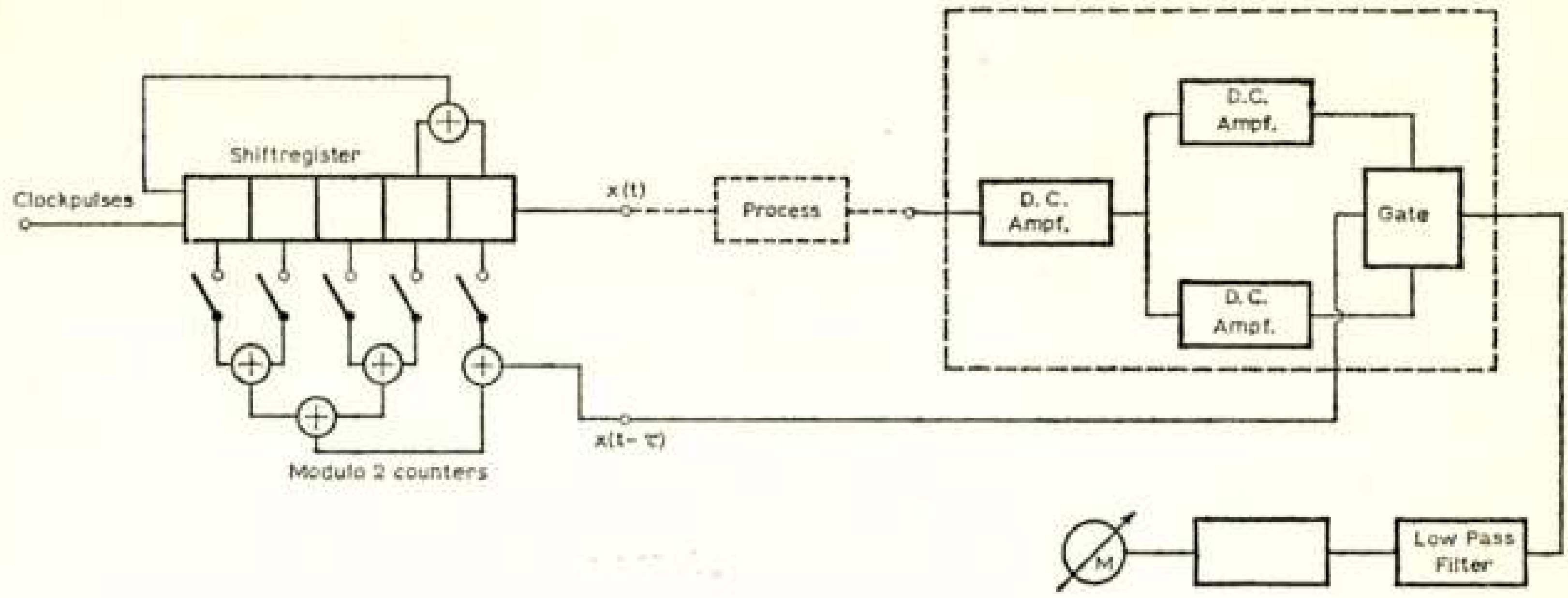

Fig. 11
Block diagram of the correlator

does not generate noise or interference, so that theoretically there is no limit to the accuracy of the measurements. In this chapter we will discuss the influence of noise on the measurements in order to determine the sensitivity of the correlator to external disturbances.

These disturbances may de added by the process itself or may be due to the input signal of the process under normal opera-

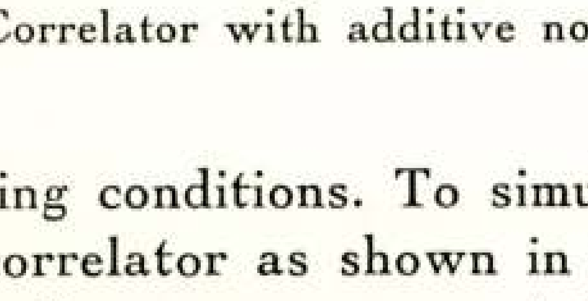

Fig. 12
Correlator with additive noise

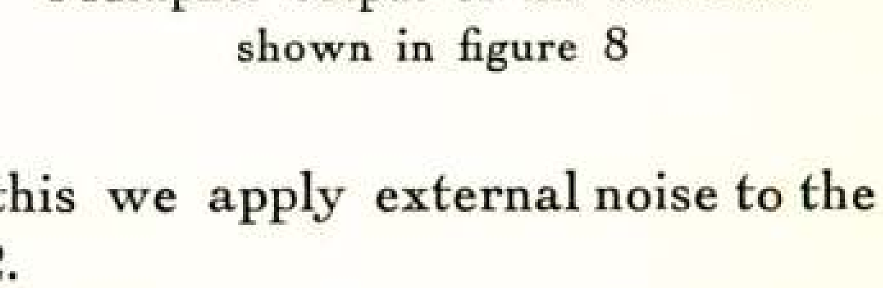

Fig. 13
Multiplier output of the correlator shown in figure 8

ting conditions. To simulate this we apply external noise to the correlator as shown in fig. 12.

To compare non periodic and periodic excitation we will discuss the following cases.

1. Excitation with the random telegraph signal viz. a periodic binary signal and a low pass filter as a mean value operator.
2. Excitation with a periodic binary signal and a periodic integrator as a mean value operator.

4.2. *Variance of $\psi_{xy}(\tau)$ using a lowpass filter as a mean value operator*

Consider now fig. 13. The convolution integral gives for the output of the lowpass filter:

$$u(t) = \int_0^\infty \{y(t-\lambda)\, x(t-\lambda-\tau) + x(t-\lambda-\tau)\, n(t-\lambda)\} g(\lambda)\, d\lambda \quad (19)$$

where $g(\lambda)$ is the impulse response of the lowpass filter.

If one assumes that the mean value of $n(t)$ equals zero and that no correlation exists between $x(t)$ and $n(t)$ then the mean value of $u(t)$ equals:

$$\overline{u(t)} = \psi_{xy}(\tau)\, Z(0)$$

$$\text{so } \psi_{xy}(\tau) = \frac{\overline{u(t)}}{Z(0)} \quad (20)$$

The variance is determined from the expression:

$$\sigma^2 = E\left[u(t) - \overline{u(t)}\right]^2 = E\left[u(t)^2\right] - \overline{u(t)}^2$$

Substitution of equation (19) leads to:

$$E\left[u(t)^2\right] = E\left[\int_0^\infty \{y(t-\lambda)\, x(t-\lambda-\tau)\, g(\lambda)\, d\lambda\right]^2 +$$

$$+ 2E\left[\int_0^\infty y(t-\lambda)\, x(t-\lambda-\tau)\, g(\lambda)\, d\lambda \,.\, \int_0^\infty x(t-\lambda-\tau)\,.\, n(t-\lambda)\, g(\lambda)\, d\lambda\right] +$$

$$+ E\left[\int_0^\infty x(t-\lambda-\tau)\, n(t-\lambda)\, g(\lambda)\, d\lambda\right]^2 \quad (21)$$

The first integral in equation (21) can be written as:

$$I_1 = E\left[\int_0^\infty y(t-\lambda)\, x(t-\lambda-\tau)\, g(\lambda)\, d\lambda\right]^2 =$$

$$= \iint_0^\infty E\left[x(t-\lambda-\tau)\, x(t-u-\tau)\, y(t-\lambda)\, y(t-u)\right] g(\lambda)\, g(u)\, du\, d\lambda$$

McFadden and G. R. Cooper [6] [7]) have shown that for the random telegraph signal the following is true:

$$E\left[x(t-\lambda-\tau)\, x(t-u-\tau)\, y(t-\lambda)\, y(t-u)\right] =$$

$$= \psi_{xx}(\lambda-u)\, \psi_{yy}(\lambda-u) + \psi_{xy}^2(\tau) + \psi_{xy}(\lambda-u+\tau)\, \psi_{xy}(\tau+u-\lambda)$$

From this it follows that:

$$I_1 = \psi_{xy}^2(\tau)\, Z^2(0) + \iint_0^\infty \psi_{xx}(\lambda-u)\, \psi_{yy}(\lambda-u)\, g(\lambda)\, g(u)\, d\lambda\, du +$$

$$+ \iint_0^\infty \psi_{xy}(\tau+\lambda-u)\, \psi_{xy}(\tau-\lambda+u)\, g(\lambda)\, g(u)\, d\lambda\, du$$

The second integral in equation (21) vanishes because $\overline{n(t)} = 0$ and $n(t)$ is not correlated with $x(t)$.

The third integral in equation (21) can be written as:

$$I_3 = \iint_0^\infty \psi_{xx}(\lambda - u)\, \psi_{nn}(\lambda - u)\, g(\lambda)\, g(u)\, d\lambda\, du$$

Substitution of these results for I_1, I_2 and I_3 results in:

$$\sigma_y^2 = \iint_0^\infty \{\psi_{xx}(\lambda - u)\, \psi_{nn}(\lambda - u) + \psi_{xx}(\lambda - u) \cdot \psi_{yy}(\lambda - u) + \\ + \psi_{xy}(\tau + \lambda - u) \cdot \psi_{xy}(\tau - \lambda + u)\}\, g(\lambda)\, g(u)\, d\lambda\, du \qquad (22)$$

This integral can be divided into two parts; one part that yields the variance due to the „finite integration time"

$$\sigma_y^2 = \iint_0^\infty \{\psi_{xx}(\lambda - u) \cdot \psi_{yy}(\lambda - u) + \\ + \psi_{xy}(\tau + \lambda - u) \cdot \psi_{xy}(\tau - \lambda + u)\}\, g(\lambda)\, g(u)\, d\lambda\, du \qquad (23)$$

and a second part that yields the variance due to the additive noise

$$\sigma_n^2 = \iint^\infty \psi_{xx}(\lambda - u)\, \psi_{nn}(\lambda - u)\, g(\lambda)\, g(u)\, d\lambda\, du \qquad (24)$$

Replacing λ en u by new variables $p = \lambda - u$ and $q = \lambda + u$ results in:

$$\sigma^2 = \frac{1}{2} \int_{-\infty}^{+\infty} \Bigg[\Big\{ \psi_{xx}(p)\, \psi_{yy}(p) + \psi_{xy}(\tau + p)\, \psi_{xy}(\tau - p) + \\ + \psi_{xx}(p)\, \psi_{nn}(p) \Big\} \int_p^\infty g\left(\frac{q - p}{2}\right) g\left(\frac{q + p}{2}\right) dq \Bigg] dp \qquad (25)$$

In order to extract the d.c. component from the multiplier output $u(t)$ it will be necessary for the smoothing filter to have a bandwidth which is small compared to that of $y(t)$ or $x(t)$.

Choosing a first order lowpass filter with a time constant T an a bandwidth smaller than the bandwidth of $y(t)$, we can write for equation (25):

$$\sigma^2 = \frac{1}{2T}\int_{-\infty}^{+\infty}\Big\{\psi_{xx}(p)\,\psi_{yy}(p) + \psi_{xy}(\tau + p)\,.\,\psi_{xy}(\tau - p) + $$

$$+\,\psi_{xx}(p)\,\psi_{nn}(p)\Big\}\,dp \qquad (26)$$

It can be shown [7]) that

$$\int_{-\infty}^{+\infty}\psi_{xx}(p)\,\psi_{yy}(p)\,dp = \int_{-\infty}^{+\infty}\psi_{xy}^2(p)\,dp$$

Using the Schwartz inequality it is easily seen that

$$\int_{-\infty}^{+\infty}\psi_{xy}(\tau - p)\,\psi_{xy}(\tau + p) \leqq \int_{-\infty}^{+\infty}\psi_{xy}^2(p)\,dp$$

Substituting these results in equation (26) leads to:

$$\sigma^2 \leqq \frac{1}{2T}\int_{-\infty}^{+\infty}\Big\{2\,\psi_{xy}^2(p) + \psi_{xx}(p)\,\psi_{nn}(p)\Big\}\,dp \qquad (27)$$

The smoothing time of a lowpass filter is defined as the integration time of a finite-memory integrator which produces the same smoothing as the actual filter.

So, if the filter has a weighting function $h(t)$, then

$$T_s = \Big[\int_0^{\infty} h^2(t)\,dt\Big]^{-1} \qquad (28)$$

For the single section RC filter with time constant $T = RC$ equation (28) results in

$$T_s = 2\,RC = 2\,T$$

Defining the correlator output signal-to-noise ratio as

$$z_0 = \frac{\psi_{xy}^2(\tau)}{\sigma^2}$$

it is now possible to express the smoothing time T_s as

$$T_s \leqq z_0\int_{-\infty}^{+\infty}\Big\{2\,\frac{\psi_{xy}^2(p)}{\psi_{xy}^2(\tau)} + \frac{\psi_{xx}(p)\,.\,\psi_{nn}(p)}{\psi_{xy}^2(\tau)}\Big\}\,dp \qquad (29)$$

Using periodic correlator input signals with period time T_1 and choosing the smoothing time T_p much greater than T_1, then the first term in equation (27) vanishes.

Hence the variance in $\psi_{xy}(\tau)$ equals

$$\sigma^2 = \frac{1}{T_p}\int_{-\infty}^{+\infty} \psi_{nn}(p)\,\psi_{xx}(p)\,dp \tag{30}$$

where $T_p >> T_1$.

The smoothing time can then be written as:

$$T_p = z_o' \int_{-\infty}^{+\infty} \frac{\psi_{xx}(p)\,\psi_{nn}(p)}{\psi_{xy}^2(\tau)}\,dp \tag{31}$$

Comparison of T_s and T_p with equal filter output signal-to-noise ratio z_o results in:

$$\frac{T_s}{T_p} = 1 + 2\,\frac{\int_{-\infty}^{+\infty} \psi_{xy}^2(p)\,dp}{\int_{-\infty}^{+\infty} \psi_{xx}(p)\,\psi_{nn}(p)\,dp} \tag{32}$$

A more compact expression can be obtained by writing (32) in terms of the ratio of mean square excitation to mean square external disturbance at the output of the system being measured.

This ratio which is:

$$z_1 = \frac{\int_{-\infty}^{+\infty} \psi_{xy}^2(p)\,dp}{\int_{-\infty}^{+\infty} \psi_{xx}(p)\,\psi_{nn}(p)\,dp} \tag{33}$$

gives a direct comparison of the relative importance of excitation and external disturbance on the system response.

In terms of this ratio (32) becomes:

$$\frac{T_s}{T_p} = 1 + 2\,z_1 \tag{34}$$

So the periodic test signal provides a smoothing which is smaller than with the random telegraph signal.

Even for z_1 as low as unity, this is a factor 3 and becomes even greater as z_1 increases. It is apparent from this that the use of periodic binary test signals offer considerable advantage.

4.3 *Variance if $\psi_{xy}(\tau)$ using a periodic integrator as a mean value operator*

The multiplier output signal is:

$$y(t)\,x(t-\tau) + x(t-\tau)\,n(t) \qquad \text{(fig. 13)}$$

So the integrator output equals:

$$u(t) = \int_0^{mT} y(t)\, x(t-\tau)\, dt + \int_0^{mT} x(t-\tau)\, n(t)\, dt =$$

$$= mT\, \psi_{xy}(\tau) + \int_0^{mT} x(t-\tau)\, n(t)\, dt$$

Under the assumption that $x(t)$ and $n(t)$ are statistically independent from each other and $\overline{n(t)} = 0$, the expectation of the integrator output is:

$$\overline{u(t)} = mT\, \psi_{xy}(\tau) \tag{35}$$

The variance is determined from the expression

$$\sigma^2 = E\left[u(t)^2\right] - \overline{u(t)}^2 = E\left[\int_0^{mT} x(t-\tau)\, n(t)\, dt\right]^2$$

or

$$\sigma^2 = \iint_0^{mT} E\left[\, x(t-\tau)\, x(\lambda-\tau)\, n(\lambda)\, n(t)\,\right] d\lambda\, dt =$$

$$= \iint_0^{mT} \psi_{xx}(t-\lambda)\, \psi_{nn}(t-\lambda)\, d\lambda\, dt$$

Replacing t and λ by the new variables $t - \lambda = p$ en $t + \lambda = q$, leads finally to

$$\sigma^2 = mT \int_0^{mT} \psi_{xx}(p)\, \psi_{nn}(p)\, dp \tag{36}$$

The integrator output signal-to-noise ratio is:

$$z_0 = \frac{\overline{u(t)}^2}{\sigma^2} = \frac{mT\, \psi_{xy}^2(\tau)}{\int_{-mT}^{mT} \psi_{xx}(p)\, \psi_{nn}(p)\, dp} \tag{37}$$

Comparing this result with equation (31) results in:

$$\frac{z_0}{z_0'} = \frac{mT}{T_p} = \frac{\int_{-\infty}^{+\infty} \psi_{xx}(p)\, \psi_{nn}(p)\, dp}{\int_{-mT}^{+mT} \psi_{xx}(p)\, \psi_{nn}(p)\, dp} \tag{38}$$

If the bandwidth of $x(t)$ is much greater than the bandwidth of $n(t)$, we can write for equation (38):

$$\frac{z_0}{z_0'} = \frac{mT}{T_p}$$

This equation is only valid if $T_p >> T$.

In order to compare the preceding results, we distinguish two cases:

a. The multiplier output signal-to-noise ratio is much smaller than one. The smoothing times for equal correlator output variance can then be written as:

$$T_s = m(1 + 2z_1)T$$
$$T_p = mT \qquad (\sigma_s = \sigma_p = \sigma_i)$$
$$T_i = mT \qquad (m >> 1)$$

It is apparent from this that the use of periodic binary signals leads to an appreciable reduction of the smoothing time.

b. The multiplier output signal-to-noise ratio is much greater than one. The smoothing times for equal correlator output variance can now be written as:

$$T_s > m(1 + 2z_1)T$$
$$T_p > mT \qquad (\sigma_s = \sigma_p = \sigma_i)$$
$$T_i = mT \qquad (m \text{ is a small integer})$$

The smoothing time T_p can not be chosen less than a certain value T'. This follows from the fact that if T_p were less than $T' = mT$, then the periodic components in the multiplier output would yield an extra contribution to the correlator output variance. From this it is clear that the use of a periodic integrator as a mean value operator leads to the shortest smoothing time.

Generally one can say that no method of system identification using a fixed observation time can do any better than the crosscorrelation with a periodic binary test signal and a periodic integrator as a mean value operator.

Acknowledgements

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References

[1]) A. Papoulis. The Fourier integral and its applications, McGraw - Hill Book Company, Inc. New York 1962, pages 81 et seq.

[2]) Solodovnikov, V.V.
Introduction to the Statistical Dynamics of Automatic Control Systems. Dover Publ. Inc. New York 1960, pages 105 et seq.

[3]) Peterson, W. W.
Error Correcting Codes.
The M.I.T. Press and John Wiley and Sons, Inc., New York, London 1961.

[4]) Fitzpatrick, G. B.
Synthesis of Binary Ring Counters of Given Periods.
Journal of the Association for Computing Machinery, Vol. 7, Number 3, pp. 287-297, July 1960.

[5]) Golomb, S. W.
Sequences with randomness properties.
Internal report. The Glenn L. Martin Company, Baltimore 3, Maryland, June 14, 1955.

[6]) McFadden, J. A.
The Fourth Product Moment of Infinitely Clipped Noise.
I.R.E. Transactions on Information Theory. Vol. IT-4, Vol. 4, December 1958, pp. 159-162.

[7]) Cooper, G. F.
Crosscorrelation with Binary Signals.
School of Electrical Engineering, Purdue University, Lafayette, Indiana.

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Buried antennas

by H. P. Williams *)

Summary

During the last few years, the subject of buried antennas has received a good deal of attention. This article summarises the theory of such antennas and is based on a lecture given to the Benelux Section of the I.R.E. and the N.R.G. on October 25, 1962.

The problem of signal-to-noise ratio is considered and an example given of communication over a distance of 50 km.

1. Introduction

The possibility of using a horizontal wire for the reception of radio signals was implicit in the classic article of 1909 by Sommerfeld [1]. In this article he calculated the wave tilt associated with the surface wave propounded by Zenneck some seven years previously. This tilting resulted in a horizontal component of the surface wave and this component can be received with wires running either just above or just below the surface of the ground.

Practical use of this fact seems to have been made by several nations in the First World War using wires above the ground. It is possible that the first actual use of a *buried* antenna was made by an official of the Dutch P.T.T., Mr. A. S. J. Vlug, who buried a receiving antenna and demonstrated that improved reception of V.L.F. signals could be obtained from Bandung in Java. As a result the Dutch P.T.T. used buried antennas from a very early date.**)

In 1923, Beverage, Rice and Kellogg [2] published their well-known paper on the wave antenna. While not strictly a buried antenna it is in fact closely related, for it relies for its pick-up on the wave tilt. The only difference from a parameter point of view is that the effective length of the wire is within some 20% of the free-space value, whereas burying the antenna would make the wavelength along the wire some 3 or 4 times less than the free-space wavelength.

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**) See also the Correspondence by Mr. J. J. Vormer on page 291.

The case of a horizontal loop antenna was studied by Elias [3] and Sommerfeld [4]. Later van der Pol [5] showed explicitly that, with a vertical magnetic dipole at the surface, the field equations along the surface consist of the difference of two terms which are identical except for the value of the propagation constant. In one case the propagation constant is that of the air, while in the other it is that of the ground. From this start we can derive, in a somewhat heuristic manner, the main field equations for a horizontal buried antenna.

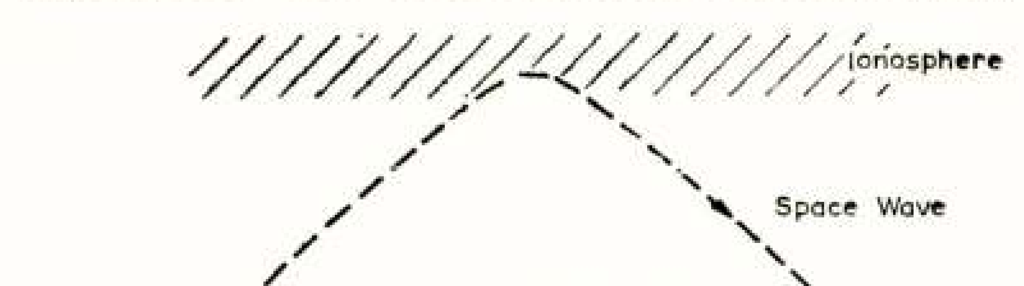


Fig. 1
The four possible paths linking two buried antennas

The propagation from a buried antenna may be divided into four distant paths as illustrated in fig. 1 which shows a buried transmitting antenna T and a buried receiving antenna R at depths d_T and d_R respectively Energy from T reaches R in four ways: two of them involve propagation through the atmosphere (and include the paths d_T and d_R), while the other two paths are wholly through the earth. An examination of the boundary conditions shows that, for propagation via the atmosphere, a buried antenna must be horizontal, for in the vertical position it is grossly inefficient.

If the distance of separation, r, is less than the sum of the depths, then only the two paths through the earth are of importance. Such a case occurs with communications between galleries down a mine and has been analyzed by Williams. [6] In this case, transmitting and receiving antennas are best placed broadside-on with respect to each other.

If, on the other hand, we have $r >> (d_T + d_R)$, then the major contribution to the field at R is via the two paths in the atmosphere. This is the more commonly used mode of propagation. In particular, it is used with small values of d_T (i.e. less than the skin depth) while d_R will either also be small or else a normal receiving antenna above the ground. A rigorous analysis of the general case is quite involved. It has been summarised by Wait [7] for the ground wave and Biggs [8] for the space wave.

In parallel with these developments, the theory of the impedance of a horizontal wire in the presence of the earth was being studied. Initially such studies were made with regard to telegraph wires, the most notable being two articles by Carson [9, 10]. Since the second World War renewed interest was aroused for military reasons and convenient formulae for the input impedance were derived by Moore [11].

A recent study is the one by Weeks and Fenwick [12] who also provide experimental confirmation for their calculations.

The analysis of buried antennas can now be said to have reached a stage where theoretical calculations can be carried out with more or less the same precision as that attainable with antennas above the ground.

2. Efficiency relative to vertical radiator

The propagation modes for the transmission from a buried antenna were outlined in the Introduction. In most applications it is the mode involving propagation via the atmosphere which is required, e.g., the case for which the distance between the antenna and the point of measurement is large in comparison with the sum of the depths of the antenna and the field point. The surface wave generated by a horizontally buried antenna can be evaluated by involving the principle of reciprocity. Suppose that in Fig. 2 the doublet antenna R is radiating, then neglecting propagation losses, the vertical electric field at Q is given by

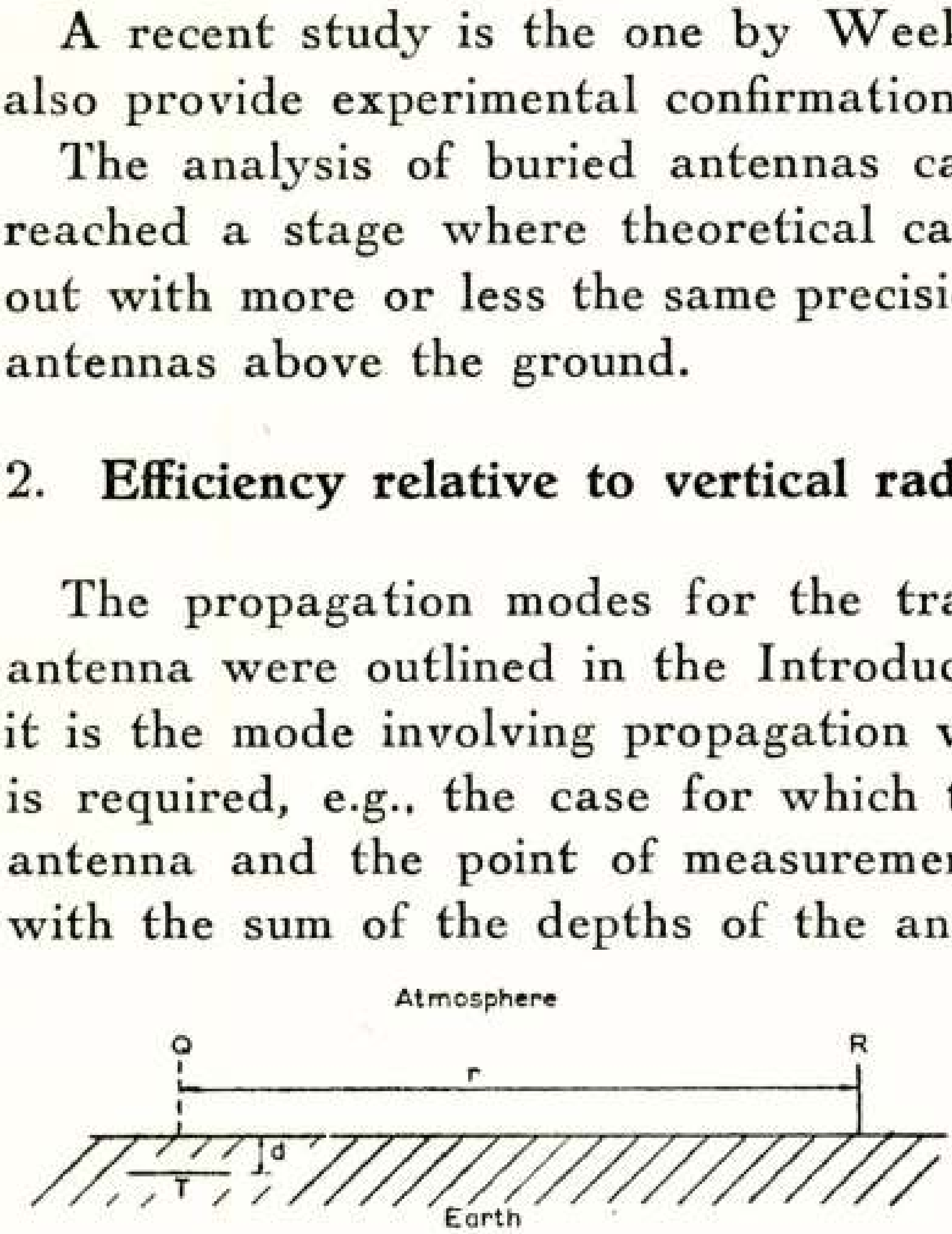


Fig. 2
Diagram for illustrating comparison between a buried and a vertical antenna

$$E_Q = z_o \frac{I\,l}{\lambda\, r} \tag{1}$$

where z_o = characteristic impedance of frese space
I = uniform current in antenna
l = length of doublet

Being a surface wave, the electric field is tilted and has a horizontal component which is $(1/n)$ of the vertical component, where n is the complex refractive index and is given by

$$n^2 = \varepsilon_r - j60\,\lambda_g \tag{2}$$

where ε_r = relative dielectric constant of ground
g = conductivity of ground in mhos/m

For wave lengths of more than 100 metres and with normal ground conductivities, the imaginary term in equation (2) dominates the real term and we have

$$|n^2| = 60\,\lambda g \tag{3}$$

Since the tangential component of E is continuous across the air-ground boundary, the horizontal field at T is given by

$$E_T = \frac{z_0}{n}\frac{I\,l}{\lambda\,r} \tag{4}$$

If the distance below the ground, d, is appreciable compared with a skin depth, then an exponential decay coefficient must be added giving

$$E_T = \frac{z_0}{n}\frac{I\,l}{\lambda\,r}e^{-\,a\,d} \tag{5}$$

where a = attenuation constant of soil
d = depth of doublet below the ground

By reciprocity this is also the vertical field at R produced by a horizontal doublet of strength I situated at T.

If the antenna at R were also a horizontal buried antenna, the field at the antenna would be $(1/n)$ of that given by (5) while burying to an appreciable extent would necessitate the addition of a further exponential decay term.

To summarize, for surface propagation the field strength due to a buried doubet is n times less than that given by a doublet of the same strength situated vertically above the ground. If the doublet is buried to a depth comparable with the skin depth, a further attenuation of d nepers must be allowed for. Typical values of a and n may be deduced from Fig. 3 where they are presented in the usually more convenient form of $1/a$ and n_2. where $1/a$ is the skin depth.

All the above equations refer to the field produced in line with the axis of the antenna. For other directions the field is proportional to the cosine of the angle made with the axis. Thus the polar pattern in azimuth is a figure of eight with

$$E = E_m \cos \emptyset \tag{6}$$

where E = vertical field in direction $\emptyset$
E_m = field in line with the antenna
$\emptyset$ = angle between line of antenna and field point

It will be noticed that this is exactly the opposite to the case of a doublet in free space, where the maxima occur at right angles to the antenna while the zeros are in line with axis of the doublet.

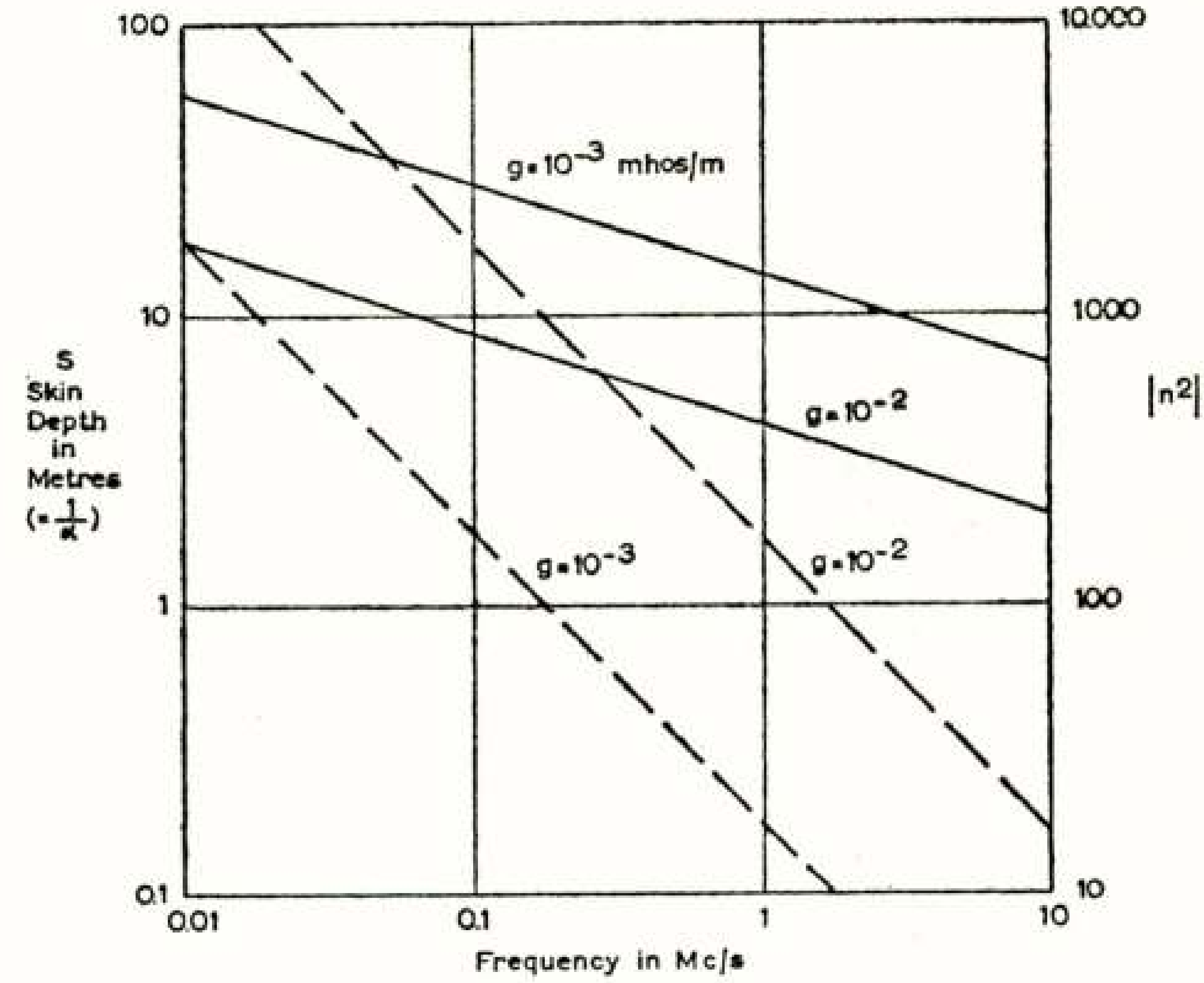


Fig. 3
Skin depth and refractive index over the range 10 kc/s-10 Mc/s

The corresponding horizontal component is equal to E/n and is radial to the source for all values of ø. Thus this component is always in the direction of propagation.

To evaluate the efficiency of a buried antenna relative to an antenna mounted vertically over the ground, we imagine a vertical doublet at Q giving the same field strength at R as the buried doublet. Equating the fields according to equations (1) and (5) gives

$$z_0 \frac{I_Q l}{\lambda r} = \frac{z_0}{n} \frac{I_T l}{\lambda r} e^{-\alpha d} \qquad (7)$$

where $I_{T,Q}$ = current in doublets T and Q respectively.

Hence the equivalent vertical doublet has a current given by

$$I_Q = \frac{I_T}{n} e^{-\alpha d} \qquad (8)$$

The relative efficiency of the buried antenna is given by

$$\eta = \frac{W_Q}{W_T} = \frac{I_Q^2 R_Q}{I_T^2 R_T} \tag{9}$$

where $R_{B,\,C}$ = resistance of doublets B and C respectively.

The value of R_Q is double that for a doublet in free space; therefore,

$$R_Q = \tfrac{4}{3}\pi z_0 \frac{l^2}{\lambda^2} \tag{10}$$

The value of R_T depends on whether the ends are shortcircuited to earth or whether the ends are left open-circuited. The appropiate input resistance in the two cases is given by equations (25) and (27) respectively. Thus we have

(a) Short-circuited horizontal wire

$$\eta_{sc} = \frac{1}{n^2} \cdot \frac{16}{3} \cdot \frac{l}{\lambda} \cdot e - 2\,a\,d \tag{11}$$

This equation is valid for $l < \lambda/20$.

(b) Open-circuited horizontal wire

$$\eta_{oc} = \frac{1}{n^2} \cdot \frac{16}{4} \cdot \frac{l}{\lambda} (e - 2\,a\,d) \tag{12}$$

This equation is valid for $l < \lambda/10$

It should be noted that equation (11) makes no allowance for the resistance to ground of the terminating electrodes. In a practical case the resistance introduced by the electrodes might well reduce the efficiency to about one half that given by (11).

As an example of the order of the efficiencies involved, let us take the case of an antenna $\lambda/20$ long and only slightly buried so that the exponential term is near unity. Then we have

$$\eta_{sc} = \frac{4}{15} \cdot \frac{1}{n^2}$$

$$\eta_{oc} = \frac{1}{5} \cdot \frac{1}{n^2}$$

Relativeto the short-circuited case the value for an open-circuit wire is increased by a factor of three on account of being one-third of the resistance but has a fourfold decrease on account of the linear current distribution. In practice the efficiency of an open-circuit wire may well be higher than the corresponding short-circuited case, since there are no electrode resistances to allow for. It can also be greater on account of the formula for η being valid up to lengths of about $\lambda/10$. In both cases increasing the length of wire beyond the limits indicated will give a somewhat higher efficiency (say 20 to 30%) at first, but as the half-wave resonance is approached in the short-circuited case, and the full-wave resonance condition in the open-circuited case, the efficiency drops rapidly. Thus the highest efficiency is given by an open-circuited wire which is in the region of $\lambda/10$ to $\lambda/6$ long. This also happens to be the region for the first resonance, i.e. half-wave resonance, as we can see from Fig. 7. In the neighbourhood of this resonant condition we have

$$\eta^{1}{}_{max} \doteqdot \frac{1}{2n^{2}} \tag{13}$$

where the raised suffix 1 indicates that this is for a single wire.

The efficiency of a buried antenna can be much improved by using several wires in parallel or series, or a combination of the two to suit the impedance of the generator. The following impedances are then obtained

Wires in parallel

$$R_m = \frac{R_1}{m} \tag{14}$$

Wires in series

$$R_m = mR_1 \tag{15}$$

where m = number of wires, each of resistance R_1

In either case the efficiency is proportional to m. With the parallel coupling the current is increased by the factor $\sqrt{m}$, while with the series coupling the total effective length is m times greater and this more than balances the $\sqrt{m}$ decrease in efficiency caused by the increase in resistance.

The maximum value of m which can be used is limited in two ways:

(1) The spacing between adjacent wires must be at least two skin depths. This is to avoid mutual coupling between the wires, which would invalidate the above equations.

(2) A point is reached where the addition of further wires would alter the basic figure-of-eight polar pattern. Gain

due to increased directivity should not be reckoned as an increase in efficiency. If we take as an arbitrary limit a maximum width of 0.3 λ for an array, then the field at 60° off the axis is 0.34 of the maximum - whereas for a single doublet it would be 0.5. Viewing the polar diagram as a whole, an array of this width shows little difference with the single doublet case.

Using the skin depth equation, we find that the above stipulations result in a maximum permissible number of elements across the width of the array which is given by

$$m_w = 0.3\,\lambda \cdot \frac{\pi}{\lambda} \cdot \frac{n}{\sqrt{2}}$$

$$= \frac{2}{3}\,n \tag{16}$$

The maximum number of elements in line clearly depends on the length of each element. The formula given in equation (13) is for an open-circuit wire of length somewhat greater than 0.1 λ. Since the wire will most probably have its half-wave resonance in the region 0.1 to 0.2 λ, this means some two to three resonant wires could be put in line. Calling this number m_L we find that the maximum number of elements in an array is given by

$$m_{max} = m_W\, m_L \tag{17}$$

$$\text{i.e.} \quad m_{max} > \frac{4}{3}\,n < 2\,n$$

No exact figure can be given for the general case. To be on the safe side we can say

$$m_{max} = n \tag{18}$$

This represents a considerable improvement in efficiency; particularly so at low frequencies where n is large, though in this case the number of elements and the total area required would also be very large. Clearly the costs and siting difficulties in such a case would have to be compared with that of a large vertical radiator with its accompanying earth network.

Multiplying equation (18) by (13) gives the best possible efficiency of an array of m elements as

$$M_{max} = \frac{1}{2n} \tag{19}$$

This is about the best that can be done without actually introducing extra directivity into the problem. To take some illustrative figures, if $f = 500\ kc/s$ and $g = 0.01\ mhos/m$, then n is about 20, so that we could expect a relative efficiency of $2\frac{1}{2}\%$ with an array of 20 elements. On the face of it this is not a very exciting prospect, but it could have applications in special cases.

3. **Radiation pattern in the air**

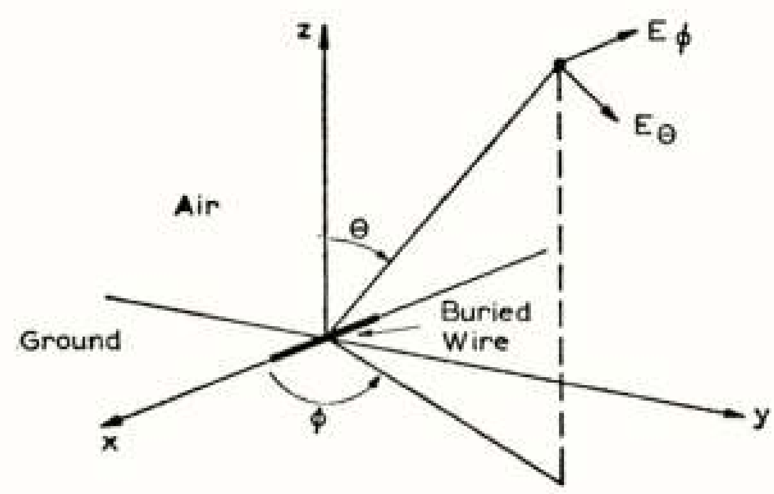


Fig. 4
Coordinate system for fig. 5

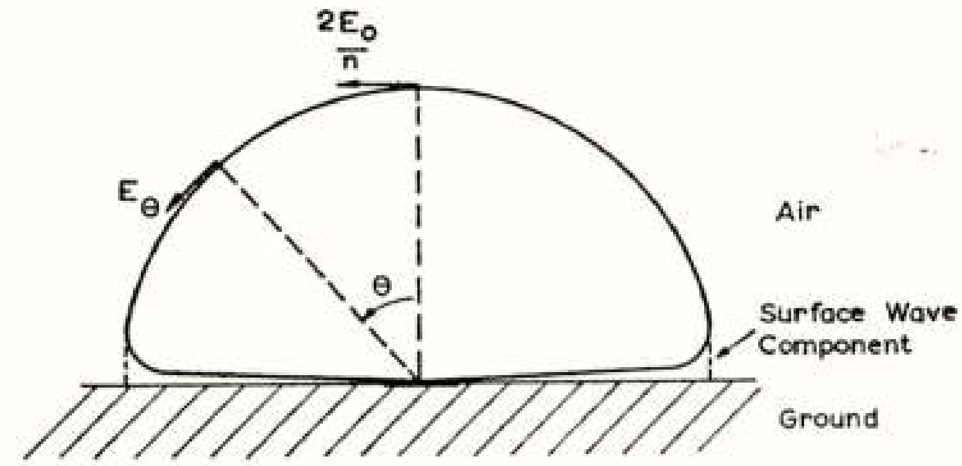


Fig. 5a
Space wave radiation patterns of a short buried antenna. In the plane $\varphi = 0°$

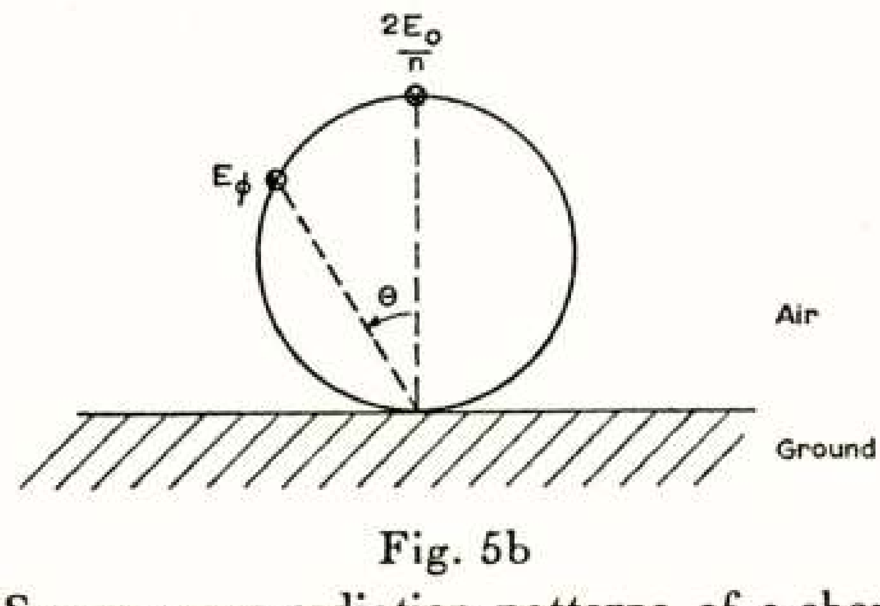


Fig. 5b
Space wave radiation patterns of a short buried antenna. In the plane $\varphi = 90°$

The radiation pattern from a buried horizontal doublet has been calculated by Biggs (8). At distances which are large in comparison with n^2 $(\lambda/2\pi)$ the results are quite simple and can be deduced, in fact, from elementary considerations of the transmission coefficient into the ground. Subject to this limitation the polar patterns in space using the coordinate system shown in Fig. 4 are as shown in Fig. 5.

The field strengths in Fig. 4 are expressed in terms of E_o, where E_o is the free space radiation field in the equatorial plane of a doublet of the same strength. This field is therefore half that given by equation (1), that is

$$E_o = z_o \frac{I l}{2 \lambda r} = 60 \pi \frac{I l}{\lambda r} \qquad (20)$$

Fig. 5a shows the pattern in the plane $\phi = 0$. For other values of ϕ we have

$$E_{\phi} = \frac{2E_0}{n} \frac{\cos\Theta}{\cos\Theta + 1/n} \cos\phi \tag{21}$$

This equation gives only the space wave. The surface wave has been considered separately at the beginning of this section when evaluating the relative efficiency of the antenna.

The pattern shown for the $\phi = 90°$ plane in Fig. 5(b) assumes that n^2 is large. For general values of ϕ we have

$$E_{\phi} = \frac{2E_0}{n} \cos\Theta \sin\phi \tag{22}$$

If there were no surface wave losses, the polar diagram in the $\phi = 0°$ plane would be a semi-circle. In this case the overall radiation pattern in the air would be half a torus whose axis is at right angles to the doublet. *Thus the general radiation pattern has the same shape as that of a doublet in free space, but is rotated through 90° and diminished in magnitude by 2/n.*

4. Input impedance

A deeply buried wire surrounded by an insulator can be regarded as a coaxial transmission line with a lossy outer conductor. Because of the lossiness, the usual *TEM* mode will not exist but will be replaced by low-order *TM* modes. The remarkable thing is that the impedance of a horizontal wire remains virtually constant as the wire is raised to the surface.

This is supported by theoretical calculations as performed, for example, by Carson [10] and Moore [11].

Experimental results by Weeks and Fenwick [12] indicate only small percentage changes as a wire is raised up to the surface. On raising the wire above the surface the resistive component only varies slowly if the wire is short (i.e. below resonance). In fact the wire must be raised to a height of about half a skin depth to cause a 10% change. However, the reactance varies much more abruptly, and a 10% change could occur as the height is increased from zero to only one-fiftieth of a skin depth.

Fig. 6 shows a cross-section of a completely buried antenna. The inner conductor of radius a_2 is assumed to be a perfect conductor, while a_1 is the inner radius of the outer conductor which is the soil itself.

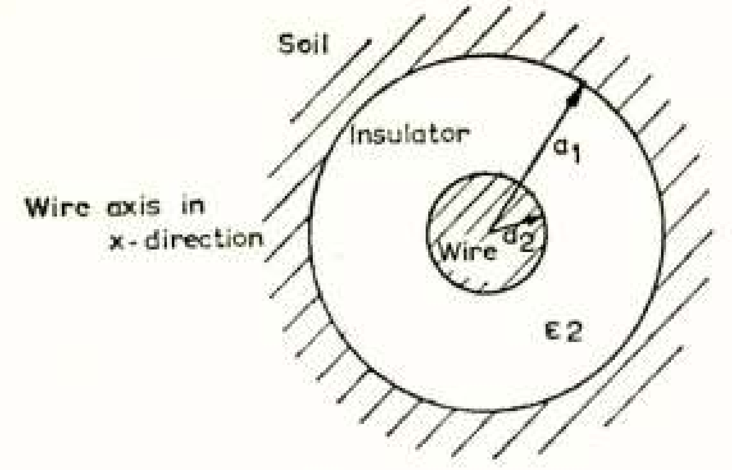


Fig. 6
Cross-section of buried wire for transmission line calculations

The equation expressing the propagation constants of the *TM* modes is in Hankel functions and cannot be solved in general except by trial and error methods using a digital computer. However, if small argument approximations can be made then we have

$$\gamma_x = \gamma_2 \left[\frac{1 - \frac{j\pi}{4} \ln 1.26 \frac{a_1}{s}}{\ln \frac{a_1}{a_2}} \right]^{\frac{1}{2}} \tag{23}$$

where γ_x = propagation constant of lossy coaxial line.

γ_2 = propagation constant of dielectric

$= -\omega^2 \mu \varepsilon$ (if lossless)

Equation (23) is under the usual assumption that n^2 is large. In addition it is assumed that $(\gamma_x^2 - \gamma_1^2)^{\frac{1}{2}} a_1$ and $(\gamma_x^2 - \gamma_2^2) a_2$ are sufficiently small for second order Hankel functions to be replaced by small arguments approximations.

The latter conditions will hold if a_1/s is less than about 0.1. A further condition is that a_1/a_2 must not be too large.

Typically the phase constant is about twice that of dielectric surrounding the wire. Since the latter will usually have a phase constant about twice that of free space, we can infer that a typical buried wire with a dielectric sheath will have a phase constant of the order of 4 times the free space value. Thus insulated buried wires have resonant lengths roughly 4 times shorter than the corresponding free space resonant length.

The characteristic impedance of the buried wire is given by

$$Z_x = \left(\frac{1}{2\pi} \ln \frac{a_1}{a_2}\right) \frac{\gamma_x}{j\omega_2} \tag{24}$$

Separating the real and imaginary parts of the input impedance, gives

(a) *wire short-circuited to ground*

$$R_{sc} = \frac{60}{\sqrt{\varepsilon_r}} \ln \frac{a_1}{a_2} \left[\frac{\beta_x}{\beta_2} \frac{\sinh 2\, \alpha_x l + \frac{\alpha_x}{\beta_2} \sin 2\, \beta_x l}{\cosh 2\, \alpha_x l + \cos 2\, \beta_x l} \right] \tag{25}$$

$$X_{sc} = \frac{60}{\sqrt{\varepsilon_r}} \ln \frac{a_1}{a_2} \left[\frac{-\alpha_x}{\beta_2} \frac{\sinh 2\, \alpha_x l + \frac{\beta_x}{\beta_2} \sin 2\, \beta_x l}{\cosh 2\, \alpha_x l + \cos 2\, \beta_x l} \right] \tag{26}$$

(b) *wire open-circuited*

$$R_{oc} = \frac{60}{\sqrt{\varepsilon_r}} \ln \frac{a_1}{a_2} \left[\frac{\beta_x}{\beta_2} \frac{\sinh 2\, \alpha_x l - \frac{\alpha_x}{\beta_2} \sin 2\, \beta_x l}{\cosh 2\, \alpha_x l - \cos 2\, \beta_x l} \right] \tag{27}$$

$$X_{oc} = \frac{60}{\sqrt{\varepsilon_r}} \ln \frac{a_1}{a_2} \left[\frac{\alpha_x}{\beta_2} \frac{\sinh 2\, \alpha_x l - \frac{\beta_x}{\beta_2} \sin 2\, \beta_x l}{\cosh 2\, \alpha_x l - \cos 2\, \beta_x l} \right] \tag{28}$$

In each case l is the length of wire from the feed point to the extremity. If a buried antenna is centre fed we have two impedances in series each one obtained from the above formulae on putting the parameter l in the equations equal to the *half-length* of the antenna. The same principle applies to an antenna with unequal *limbs*. Thus, if the impedance of one side is $R_1 + jX_1$ and of the other $R_2 + jX_2$ the input impedance is given by

$$R = R_1 + R_2 \tag{29}$$

$$X = X_1 + X_2 \tag{30}$$

Fig. 7 shows an impedance curve derived from the above six formulae. Excepting in the immediate neighbourhood of the resonance points, experimental results by Weekes and Fenwick [12] gave good agreement with the above theory and showed that it is valid even in the case of unbalanced antennas.

The resonance conditions are given by putting $X = 0$ in equations (26) and (28) and are therefore

$$\pm \sin 2\, \beta_x l = \frac{\alpha_x}{\beta_x} \sinh 2\, \alpha_x l \tag{31}$$

The plus sign gives the short circuit resonances, while the minus sign gives the open circuit resonances.

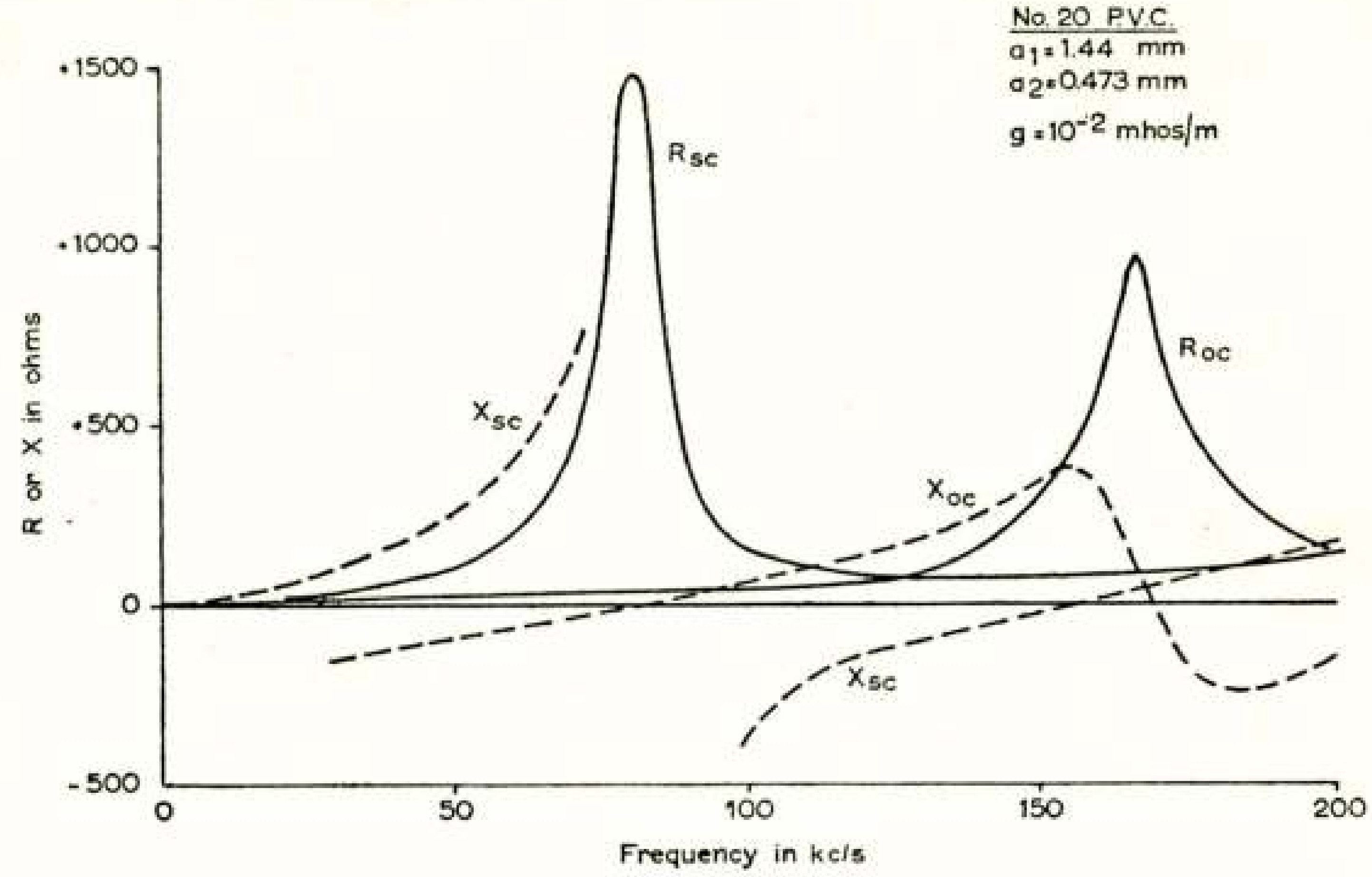


Fig. 7
Impedance of 450-m buried wire

It can also be shown from the lossy line theory that the current distribution is approximately sinusoidal, provided $a_x l$ is small. In particular for short antennas the current is either

(a) uniform — for wires with grounded ends, i.e. short-circuited wires

or (b) linear — for wires with insulated ends, i.e. open-circuited wires

The impedance equations for short buried wires are particularly simple. By "short" we mean wires in which the current is either substantially uniform (wire with grounded ends) or substantially linear (wire with insulated ends). The approximate limits for these conditions are given by

(a) *short-circuited wire*

$$l_{sc} < \frac{\textit{half-wave resonant length}}{4} \qquad (32)$$

(b) *open-circuited wire*

$$l_{oc} < \frac{\textit{half-wave resonant length}}{2} \qquad (33)$$

For most practical cases these limits correspond roughly to $l_{sc} < \lambda/20$ and $l_{oc} < \lambda/10$.

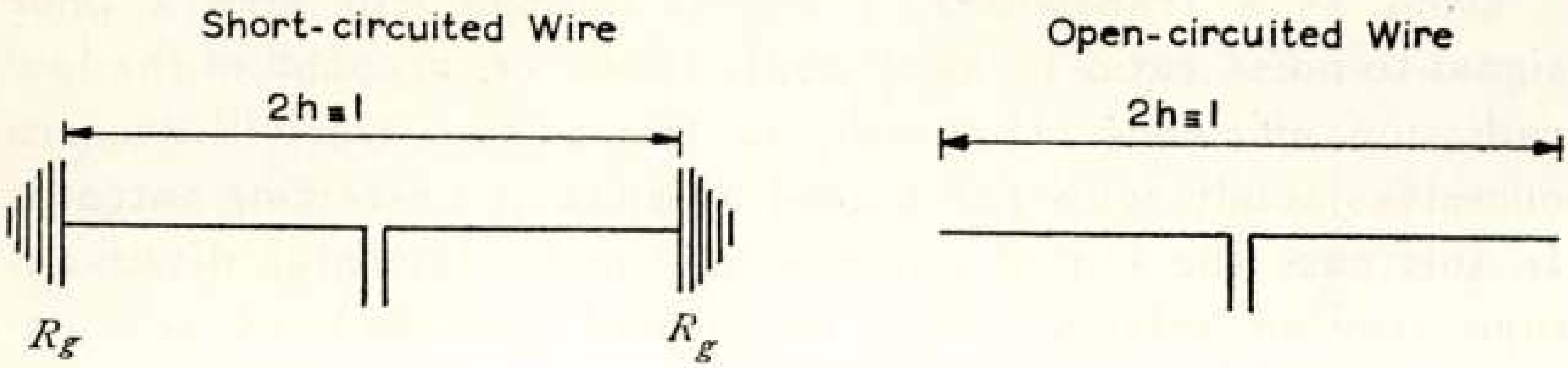


Fig. 8
Sketches of grounded and insulated buried horizontal wires

Under these conditions we have (see Fig. 8).

(a) *Grounded ends*

$$R_{sc} = 30\,\pi^2\, l/\lambda \tag{34}$$

$$X_{sc} = \frac{4\,R_{sc}}{\pi}\, ln\, \frac{0.794\,s}{a_2} \tag{35}$$

The resistance and reactance are the same for all positions of feed-point since the current is uniform.

In practice the total resistance will be greater because of the additional resistance due to the two grounding electrodes. Formulae for such electrodes are given in [6]. The following is for a sunken rod whose top is level with the surface

$$R_g = \frac{1}{2\,\pi\, g\, l_r}\, ln\, \frac{2\, l_r}{a} \tag{36}$$

where l_r = length of rod
a = radius of rod

(b) *Insulated ends*

$$R_{oc} = 10\,\pi^2\, l/\lambda \tag{37}$$

$$X_{oc} = -\,\frac{1200}{\varepsilon_2\, R_{oc}}\, ln\, \frac{a_1}{a_2} \tag{38}$$

Here again the resistance is independent of the position of the feedpoint. But this is not so for the reactance in view of the inverse relationship with R_{oc}. In particular, an end-fed wire will have one-quarter of the reactance of a centre-fed wire of the same overall length.

5. Signal-to-noise considerations

Used as a transmitter, a buried antenna will give a poor signal-to-noise ratio in most applications on account of the low radiation efficiency. However, in this section we will concern ourselves solely with the buried antenna as a receiving antenna. In this case the buried antenna has no fundamental disadvantage over an antenna above the ground since the extra attenuation suffered by the incoming signal also applies to the incoming noise. Over the range of frequencies for which buried antennas are used, the noise is primarily atmospheric noise and therefore all the significant noise suffers the same attenuation as the signal. The same is even true for the VHF part of the spectrum, where the main noise is of cosmic origin.

Nevertheless, there are two factors which modify this equality of performance. One works in favour of and the other against the buried antenna. The factors are:

(1) There is a reduction of noise of the static kind. This has been noticed by a number of organisations including the Dutch PTT.

(2) There are other sources of noise in the ground itself. This factor is mainly of interest for very low frequencies in the audio band. In addition, at such low frequencies we may have noise from power lines.

The extra sources of noise in the ground vary according to location and are a study in themselves. In the present section we will assume that only atmospheric and cosmic noise are of significane.

(a) *Noise in the range* 1 *to* 100 *kc/s*

Atmospheric noise curves down to a frequency of 10 *kc/s* have been published in CCIR Report No. 65 (13). An extension to still lower frequencies can be made by using the data given by Watt and Maxwell [14]. This has been done in fig. 9 which also includes the corresponding CCIR curve. The dip at 4 *kc/s* is characteristic of propagation at this low frequency; it is due to the waveguide nature of the propagation between the earth and the ionosphere. This size of the dip is indeed indicative of the fact that most of the noise originates from sources some 1000 to 2000 km away. The reason for this feature is simply that the nearer sources are relatively fewer, while the further sources are relatively weaker.

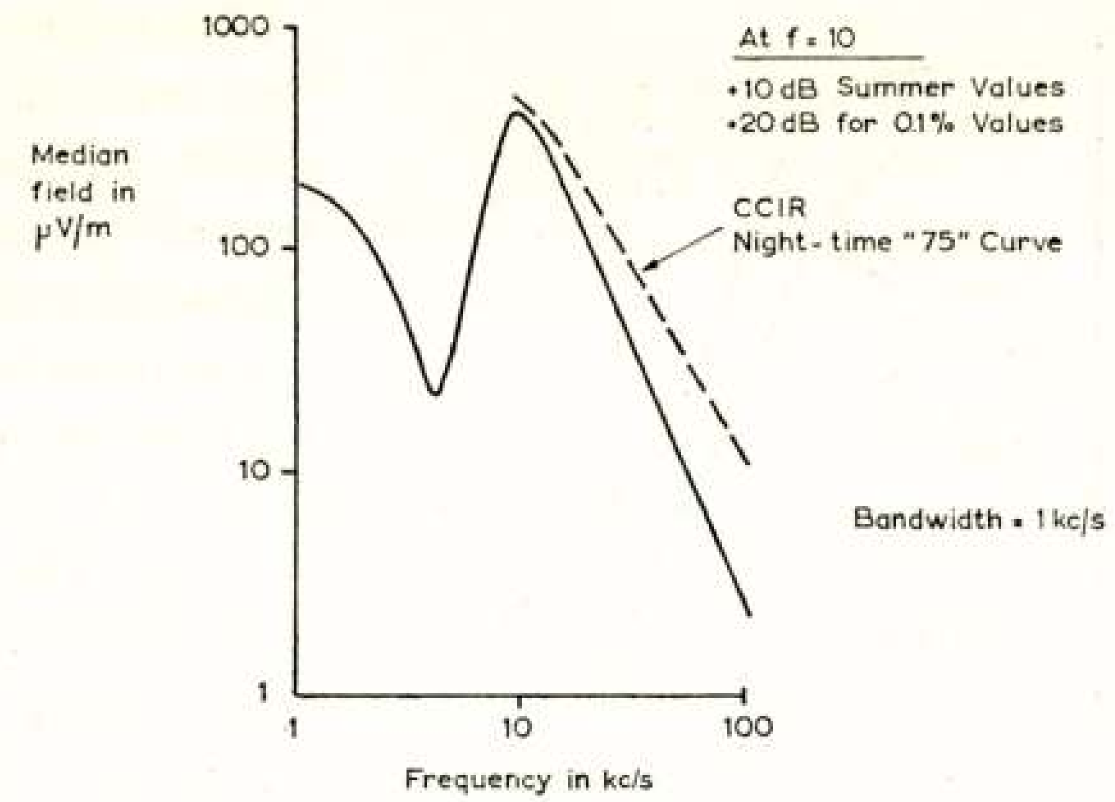


Fig. 9
Example of noise fields in the range 1-100 kc/s

Fig. 9 is for the vertical component over the ground. Translating this into the horizontal component means dividing by the refractive index, n. This has been done in Fig. 10 in which the receiving wire is assumed to be 100 metres long and grounded at both ends (i.e. the effective "height" is equal to its actual length). It will be seen that the lower frequencies give less noise. They will also pick up less signal from a surface wave by the same token.

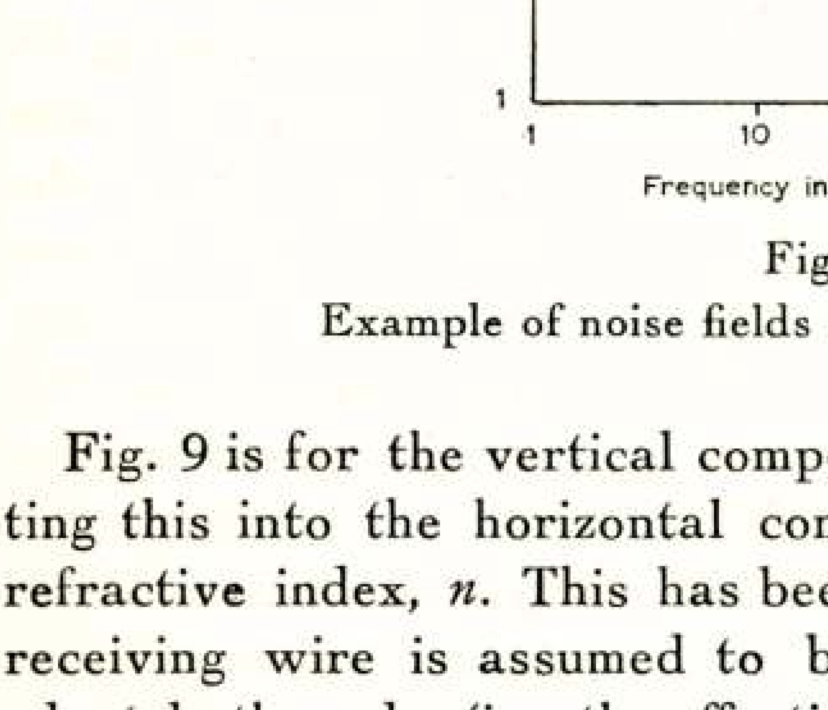


Fig. 10
Noise voltage picked up by horizontal 100-m wire if in noise field of fig. 9

Below 1 *kc/s* the noise in the ground is particularly speculative since we are running into the region of semiconductor effects.

(b) *Noise in the range* 100 *kc/s to* 100 *Mc/s*

This range is adequately covered by the CCIR curves. An example is given in Fig. 11 which shows a case of high atmospheric noise level further enhanced by taking the level exceeded by only 1% of all hourly means. On the same figure is plotted the field at 50 km distance from a short vertical antenna radiating 1 kW of power.

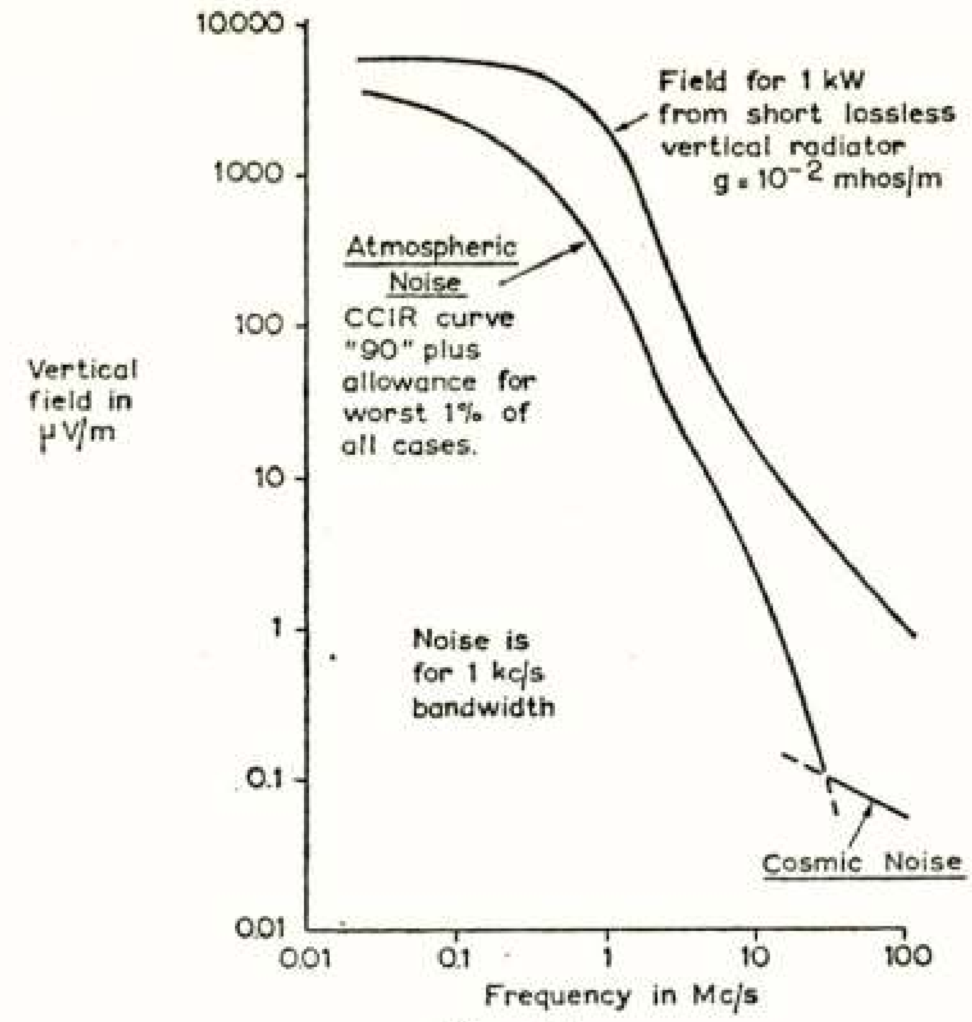


Fig. 11
Noise fields in the range 100 kc/s-100Mc/s and signal at 50 km from a short vertical radiator

From these two curves we can obtain a curve of power required to give a fixed signal-to-noise ratio for a given bandwidth (in this case 10 $dB\,S/N$ for a 1 kc/s bandwidth). This is the dotted curve in fig. 12. The result is quite suprising since the VHF range shows up very favourably with respect to MF and LF transmissions. The reason why groundwave VHF transmissions are not normally used is simply that obstacles such as buildings and trees can cause heavy additional attenuation of the signal. Furthermore, in urban

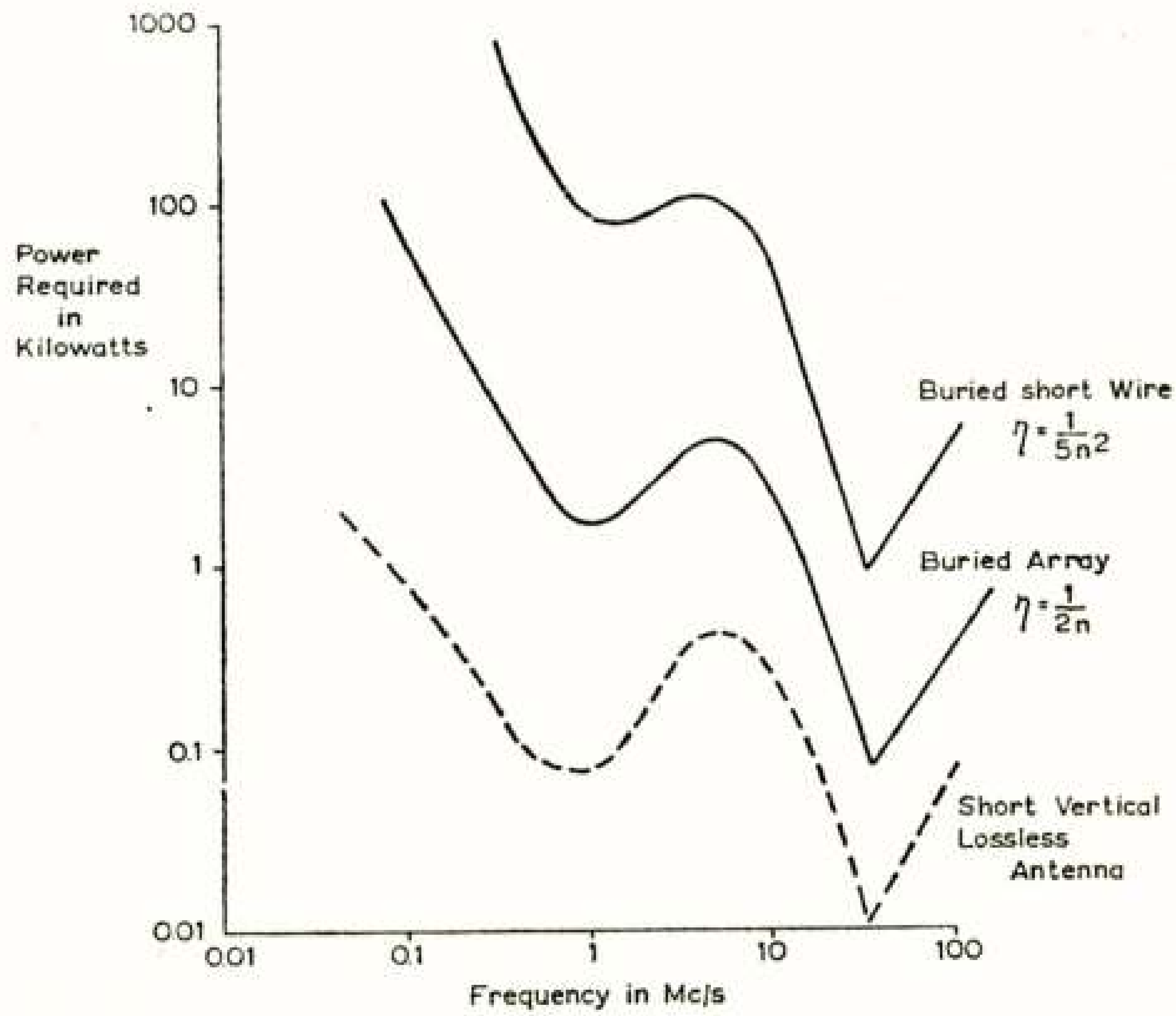


Fig. 12
Power required for 10-dB SNR at 50 km for 1-kc/s bandwidth

areas the man-made noise can be very considerable — some fifty times the e.m.f. due to cosmic noise.

Multiplying by the appropriate refractive index curve of Fig. 3 gives the full line in Fig. 12 and this distinctly favours the higher frequencies. In particular, there is a strong minimum at 30 Mc/s which is where the atmospheric noise drops very sharply to a level below the cosmic noise.

Actually it is unlikely that one would contemplate the use of VHF surface waves for the reasons given above. Therefore, VHF buried antennas are hardly likely to be used this way. It is a somewhat different matter where skywaves are concerned, for here there may well be applications when buried antennas could be used for reception, or even transmission, on the HF and even on the VHF bands.

6. Conclusions

Except for transmissions between deeply buried locations, the best performance is always obtained when a buried antenna is used horizontally. This results in a figure-of-eight polar diagram in the horizontal plane, the maximum being along the axis of the antenna. The overall polar diagram in the air has the shape of half a torus.

A useful form of buried antenna is the resonant half-wave dipole. Due to the presence of the earth, its actual length is usually between $\lambda/6$ and $\lambda/8$, while the input resistance is of the order of 20 ohms. This resistance is a loss resistance associated with the induction field: the radiation resistance is very small since the efficiency of the antenna is only about $1/2n^2$, where n is the refractive index. For example, for a frequency of 500 *kc/s* and a conductivity of 0.01 *mhos/m* the efficiency would be about 0.1%. By making an array of 20 such dipoles, the efficiency could be increased to 2.5%.

Buried antennas have the following basic advantages and disadvantages:

(1) Used as receiving antennas, they avoid a good deal of the precipitation static noise to which antennas above the ground are subject.

(2) Used as transmitting antennas they are of very low

efficiency and would not be used except for reasons of economy or military hardening.

(3) Their directional radiation pattern can be of advantage in VLF or LF operation (where the normal practical antenna is a monopole and therefore omnidirectional in azimuth). In reception, this feature has been used for the creation of a cardioid polar pattern for minimising interference. In transmission, this property permits the generation of a skywave in a given direction with a negligible surface wave.

REFERENCES

1. Sommerfeld, A., „Ueber die Ausbreitung der Wellen in der drahtlosen Telegraphic". Ann. der Phys. v. 28 p. 655 1909.
2. Beverage, Harold H., Rice, Chester W.; and Kellogg, Edward W., "The wave antenna - a new type of highly directive antenna". Trans. A.I.E.E. v. 42 p. 215 Febr. 1923.
3. Elias, G. J., „Het electromagnetisch veld van een magnetische dipool". Physica v. 2 p. 207 and p. 361 July and Dec. 1922.
4. Sommerfeld, A., „Ueber die Ausbreitung der Wellen in der drahtlosen Telegraphie". Ann. der Phys. v. 81 p. 1135 1926.
5. van der Pol, Balth., „Ueber die Ausbreitung elektromagnetischer Wellen", Jahrb. Drahtl. Tel. v. 37 p. 152, April 1931.
6. Williams, H. Paul, "Subterranean communication by electric waves", J. Brit. I.R.E. v. 11 p. 101 March 1951.
7. Wait, James R., "The electromagnetic fields of a horizontal dipole in the presence of a conducting half-space". Can. Journal of Physics v. 39 pp. 1017-1028, July 1961.
8. Biggs, Albert W., "Radiation fields from a horizontal electric dipole in a semi-infinite concucting medium". I.R.E. Trans. vol. AP. 10 pp. 358-362, July 1962.
9. Carson, J. R., "Wave propagation over parallel wires: the proximity effect", Phil. Mag. v. 41, April 1921.
10. Carson, J. R., "Wave propagation in overhead wires with ground return", B.S.T.J. v. 5, pp. 539-554, October 1926.
11. Moore, R. K., "The theory of radio communication between submerged submarines", Ph. D. dissertation, Cornell University 1951.
12. Weeks, W. L. and Fenwick, R. C. "Flush or Buried Antenna structures", Collins Radio Company, Research Report CRR-257, 1962.
13. C. C. I. R. Report No. 65. "Revision of atmospheric radio noise data", International Telecommunications Union, Geneva 1957.
14. Watt, A. D. and Maxwell E. L., "Characteristics of atmospheric noise from 1 to 100 kc". Proc. I. R. E. v. 45 pp. 787-794, June 1957.

Correspondentie „Buried Antennas"

In aansluiting op het artikel van de heer Williams, waaruit men opmaakt dat de ontvangst op ingegraven antennes blijkbaar opnieuw in de belangstelling staat, is het misschien wel aardig te vermelden, dat ontvangst door middel van ingegraven draden bij de Nederlandse PTT meer dan 40 jaar in gebruik is geweest. Juist onlangs zijn de laatste ontvanginrichtingen op het ontvangstation Nederhorst-den-Berg, die met aarddraden-kruisen uitgerust waren, opgeruimd, aangezien de Nederlandse PTT geen radioverbindingen op lange golven meer onderhoudt met andere Administraties.

Aarddraden zijn voor een ontvangcentrale, waar men vele stations gelijktijdig wenst te ontvangen, bijzonder geschikt, aangezien het daarbij mogelijk is zonder onderlinge beïnvloeding een groot aantal ontvangers op één antenne aan te sluiten.

Bij een beschrijving van het ontvangstation NORA, gepubliceerd in 1932 [1]), wordt een uitvoerige beschouwing gewijd aan gelijktijdige ontvangst van een aantal lange-golf stations op één aarddraden-kruis. Daarbij wordt gebruik gemaakt van een aantal parallel geschakelde goniometers, terwijl de energie, nodig om eenzijdige ontvangst te verkrijgen, eveneens voor alle stations van één gemeenschappelijke antenne afgeleid wordt.

De ontvangst met behulp van aarddraden was in die tijd echter een bij PTT reeds lang toegepaste techniek. De eerste mededelingen hiervoor gaan terug tot 1920 toen in „Radio Nieuws" een berichtje [2]) verscheen, waarin medegedeeld werd, dat de heer A. J. S. Vlug, een ambtenaar van PTT die zich intensief met proeven op dit gebied bezig hield, op deze wijze van ontvangst octrooi had aangevraagd.

's-Gravenhage 26-4-1963 J. J. Vormer

[1]) „Het ontvangstation Noordwijk-Radio", door Dr. Ir. N. Koomans; de Ingenieur 13 mei 1932, nr. 20, pag. E 43 e.v.

[2]) „Ontvangsysteem A. S. J. Vlug" door C(orver). Radio Nieuws 1920 pag. 6.

Xe PLENAIRE VERGADERING VAN HET CCIR TE GENEVE (15 JANUARI—15 FEBRUARI 1963)

1. Samenstelling van de Nederlandse Delegatie.

Hoofd	Ir. A. D. J. Uurbanus
Leden	
Studiegroep I	: Ir. M. C. Ennen
„ „ II	: Ir. K. Vredenbregt
„ „ III	: Dr. Ir. H. C. A. van Duuren
	Ir. K. Vredenbregt
„ „ IV	: Ir. J. J. Vormer
	Ir. L. Krul
„ „ V en VI	: Ir. J. C. Dito
„ „ IX	: Ir. J. J. Vormer
	Ir. L. Krul
	Ir. F. R. Neubauer
„ „ X	: Ir. J. C. Verton
	Ir. P. H. Boukema
„ „ XI	: Ir. F. Maarleveld
	Ir. J. C. Verton
	Ir. P. H. Boukema
„ „ XIII	: D. J. van Doorninck
CMTT	: Ir. G. Radstake
Expert	Dr. Ir. J. J. Geluk

2. Beknopt verslag van de werkzaamheden in de verschillende studiegroepen.

STUDIEGROEP I *(zenders)*.

De problemen in studiegroep I omvatten samenvattend de technische eigenschappen van zenders zoals frequentietolerantie, vermogen bij verschillende modulatiesystemen alsmede de met deze zenders verbonden gewenste en ongewenste uitzendingen. Zo werden de recommandaties en studieprogramma's over spectrum en bandbreedte van uitzendingen bewerkt en aangevuld. Voor wat betreft de frequentietoleranties zij een rapport genoemd waarin een tabel is opgenomen die voor een aantal stationstypen op verschillende frequenties aangeeft welke toleranties momenteel of in de naaste toekomst bereikbaar zijn, terwijl tevens de volgens het CCIR uiteindelijk bereikbare toleranties zijn aangegeven.

Er werd een nieuwe recommandatie ontworpen over de relatie tussen de verschillende voor een zender op te geven vermogens. De bedoeling van deze recommandatie is een aantal definities te geven met betrekking tot het gedrag van radiozenders. Met name laat men het vermogen van de zender afhangen van de optredende vervorming. Een en ander wordt door aanbevolen meetmethoden vastgesteld. Ook bevat de recommandatie een tabel van omrekeningsfactoren waarmede het mogelijk is het vermogen van een zender (belast met „smoothly-read-text") uit te drukken in piekvermogen, draaggolfvermogen en gemiddeld vermogen naar keuze.

Andere onderwerpen die ter sprake kwamen in deze studiegroep waren de industriële storingen en de classificatie van uitzendingen door middel van een code.

STUDIEGROEP II *(ontvangers)*.

De aanbevelingen betreffende gevoeligheid, selectiviteit en stabiliteit ondergingen een redactionele wijziging en werden in de tabellen met nieuwe meetgegevens aangevuld. Het ligt in de bedoeling, deze tabellen in de toekomst geleidelijk te doen vervallen. Daarvoor in de plaats zullen lijsten worden gepubliceerd met waarden die karakteristiek zijn voor de kenmerkende grootheden van een gemiddelde ontvanger, voor een bepaald welomschreven doel ontworpen. Deze lijsten stellen „Typical receivers" voor, dat wil zeggen, zij geven de eigenschappen van de gemiddelde ontvanger, ontworpen voor het beoogde doel volgens de stand van het hedendaagse technische kunnen. Voor het opstellen van deze lijsten is een studieprogramma voor „typical receivers" in het leven geroepen. Het werd gewenst geacht dat ieder land de reeds ontworpen lijsten in eigen kring nader onderzoekt.

In verband met de nieuwe mogelijkheden voor de stabiliteit van ontvangers werd de vraag omtrent stabiliteit van de afstemming omgewerkt. Ook de recommandatie en het rapport betreffende dit onderwerp ondergingen wijzigingen.

De bestudering van een vraag over de keuze van de middenfrequentie van superheterodyne ontvangers werd afgesloten, nadat het desbetreffend rapport was aangevuld met een paragraaf omtrent televisie- en scheepsontvangers.

Het rapport dat handelt over de straling van ontvangers werd uitgebreid met stralingsgegevens van televisie-ontvangers voor u.h.f en v.h.f.

Na de tussentijdse vergadering van de studiegroep in 1962 was per correspondentie een vraag geaccepteerd die op diversity-ontvangst betrekking heeft, en die in eerste opzet door de USSR was ingediend.

STUDIEGROEP III *(systemen voor vaste diensten).*

Op het gebied van telefoniesystemen werd vooral aandacht besteed aan enkelzijbandsystemen. De kanaalligging in een enkelzijbanduitzending werd nader gepreciseerd. Deze is nu uniform voor het gehele frequentiegebied beneden 30 MHz. Een vraag werd geformuleerd betreffende de mogelijkheid om bij een telefonieverbinding in beide richtingen van dezelfde radiofrequentieband gebruik te maken.

De telegrafiesystemen nemen een belangrijke plaats in onder de radiotelecommunicatiemogelijkheden voor de toekomst. Dit bleek onder andere uit de behoefte om de onderverdeling van een enkelzijbandkanaal in een aantal synchrone 100 baud-telegrafiekanalen vast te leggen.

Op het gebied van facsimile en fototelegrafiesystemen werden de f.m.-karakteristieken vastgelegd voor facsimile-uitzendingen van weerkaarten.

Voor de gebruikelijke communicatiesystemen beneden 30 MHz werden de toleranties in ontvang- en zendfrequenties vastgelegd voor die gevallen, waar men aan de ontvangzijde van frequentiebijregeling wil afzien.

De inhoud van een aantal rapporten is gewijd aan de optimale breedte van telegrafiekanalen en aan de eigenschappen van telegrafiedetectoren bij ontvangst van signalen, gestoord door ruis en fading.

Aandacht werd besteed aan telegrafiesystemen met automatische navraag. Een aanbeveling en een rapport van de vorige plenaire zitting van het CCIR werden in een nieuwe aanbeveling verwerkt die nu ook afspraken bevat betreffende het automatisch in fase brengen van synchrone systemen met meer kanalen door tijdverdeling. Voor verdere ontwikkelingen op dit gebied werd een studieprogramma opgesteld.

Een uitvoerig rapport is verschenen betreffende het meewegigheidseffect op lange afstandverbindingen.

Tenslotte werd een internationale werkgroep opgericht die definities en richtlijnen zal opstellen voor het optimale gebruik van het radiofrequentiespectrum. Dit onderwerp was aanvankelijk door de USA-afvaardiging in studiegroep II ter sprake gebracht, doch werd later verwezen naar studiegroep III.

STUDIEGROEP IV *(ruimte-communicatie).*

Het hypothetisch referentiecircuit voor de aardse communicatie met behulp van satellieten zal bestaan uit één hop inclusief modulatie- en demodulatie-apparatuur, waarbij ervan wordt uitgegaan dat de lengte van één hop (gemeten langs een grote cirkel) minstens 7500 km zal bedragen. Op grond van deze aanname konden vervolgens voorlopige waarden worden aanbevolen voor de minimale transmissiekwaliteit bij multikanalen- telefonie- en televisie-overdracht.

Geconstateerd werd dat gemeenschappelijk gebruik van bepaalde frequentiebanden door aardse straalverbindingen en satellietverbindingen voor aardse communicatie onder bepaalde restricties mogelijk moet zijn. Deze restricties moeten dan zowel aan de satellietverbinding als aan de straalverbinding worden opgelegd. Deze conclusie is bijzonder belangrijk omdat dit een uitgangspunt kan zijn, voor de dit najaar te houden administratieve radioconferentie.

Navigatie-, meteorologie- en omroepsatellieten vormen het onderwerp van verdere studie.

Door studiegroep IV werd een document samengesteld waarin een overzicht wordt gegeven van de radio-astronomiewaarnemingen die worden verricht alsmede van de apparatuur die hiervoor in gebruik is.

STUDIEGROEP V *(troposferische propagatie).*

Ten behoeve van het werk in studiegroep IV werd een rapport samengesteld waarin krommen worden gegeven die van belang zijn bij het bestuderen van interferentieproblemen in het frequentiegebied van 1-10 GHz. Over deze materie werd eveneens een nieuw studieprogramma geformuleerd.

Verder werden nieuwe rapporten samengesteld over fadingverdelingen bij straalverbindingen en de constanten in de vergelijking voor de refractie-index.

STUDIEGROEP VI *(ionosferische propagatie).*

Er zijn een aantal nieuwe aspecten van het ionosfeeronderzoek. In de eerste plaats de toenemende belangstelling van scheve-invalmetingen waarbij de nadruk komt te liggen op transmissiemetingen terwijl de terugstrooimetingen op de achtergrond raken. Verder wil men trachten een betere interpolatie van gegevens te krijgen voor gebieden waar geen peilingen worden uitgevoerd door toepassing van elektronische rekenmachines. Een derde aspect is de peiling van de ionosfeer met gebruikmaking van satellieten boven het F2-maximum („topside sounding").

STUDIEGROEP VII *(standaardfrequenties en tijdseinen).*

De documenten voor standaardfrequentie-uitzendingen en tijdseinen werden omgewerkt en aangevuld. Een speciale rol speelde bij de besprekingen de vermindering van interferentiestoringen in de overlappingsgebieden. Er werd ook een grotere frequentieconstantheid $(5 . 10^{-10})$, veelvuldiger afstemming op UT2 en vergelijking met de atomaire Caesium-standaard aanbevolen.

STUDIEGROEP VIII *(controle van uitzendingen).*

In deze studiegroep houdt men zich voornamelijk bezig met de meting van frequentie, bandbreedte en veldsterkte van de te controleren uitzendingen.

Voor wat betreft de frequentiemetingen worden de toleranties voor deze metingen boven 10,5 GHz verruimd. Een nieuw studieprogramma behandelt de bandbreedtemetingen. Op het gebied der veldsterktemetingen is de vraag naar geschikte meetmethoden actueel in verband met de uitbreiding van het frequentiegebied tot 1 GHz en de verhoging van de meetnauwkeurigheid.

STUDIEGROEP IX *(straalverbindingen).*

Deze studiegroep heeft zich voor een belangrijk deel van de bijeenkomst bezig gehouden met de studie van de mogelijkheid de voor straalverbindingen bestemde frequenties ook voor de communicatiesatellieten te bestemmen.

De resultaten tijdens de interimbijeenkomst te Parijs bereikt, werden in het algemeen zonder grote wijzigingen goedgekeurd. Een aantal belangrijke punten volgen nu.

De voornamelijk in de Verenigde Staten en Japan toegepaste kanaalindeling voor de 4 GHz-band kreeg in de nieuwe versie van Recommandatie 278 een andere plaats, zonder tot aanbevolen systeem te worden verheven. Nieuwe kanaalindelingen werden vastgesteld voor de banden 6425-7125, 8200-8500 en 10700-11700 MHz.

Ook voor circuits korter dan 280 km werd nu de toelaatbare ruis vastgesteld, een nadere studie is gevraagd van de eisen die gesteld moeten worden aan circuits die qua samenstelling belangrijk van het hypothetisch referentiecircuit afwijken.

De recommandatie betreffende de gelijktijdige overdracht van een televisiesignaal en een geluidskanaal werd opnieuw geredigeerd waarbij ook het Russische systeem werd opgenomen.

STUDIEGROEP X *(omroep)*.

De in het „Chairmansreport" opgenomen stukken werden ter behandeling verdeeld over 3 subgroepen.

Subgroep X-A behandelde de problemen betreffende geluids- en videoregistratie en het meten van het niveau en de ruis. Subgroep X-B behandelde de hoogfrequente omroepproblemen (lange golf, middengolf en korte golf). Subgroep X-C nam o.a. het onderwerp stereofonie voor haar rekening.

Dit laatste onderwerp trok verreweg de grootste belangstelling. Op de interimbijeenkomst die gehouden werd in 1962 te Bad-Kreuznach was besloten om een ontwerp-recommandatie voor een in te voeren systeem voor stereofonische uitzendingen op f.m.-zenders ter behandeling voor te leggen aan de Xde Plenaire vergadering. Enkele landen, met name Engeland en Rusland behielden zich hierbij door middel van voetnoten enige reserves voor.

Bij de bespreking van dit document bleek al spoedig dat van verschillende zijden verzet kon worden verwacht tegen aanvaarding van bedoelde aanbeveling. Zweden meldde nog een nieuw systeem in ontwikkeling te hebben, van Russische zijde werd bezwaar gemaakt tegen een frequentiezwaai van 75 kHz. Ook schaarde de Engelse Administratie zich achter de tegenstanders in verband met het Engelse bezwaar tegen het verschil in verzorgingsgebied van monofonische en stereofonische uitzendingen, dat overigens inherent is aan elk systeem.

Een enthousiaste verdediging zowel door de voorzitter van Studiegroep X mr. Prose Walker als door de voorzitter van de Subgroep Dr. Geluk, konden niet verhinderen dat overeenstemming over een recommandatie uitbleef.

In de vorm van een rapport werd het stereofonieprobleem aan de plenaire zitting voorgelegd. De vertegenwoordigers van West-Duitsland, Frankrijk, Italië en Nederland verklaarden geen behoefte meer te voelen om de systeemkeuze uit te stellen en bovendien achtten zij voortzetting van proefnemingen met andere bekende systemen niet meer nodig.

In de eerstvolgende interimbijeenkomst van Studiegroep X die gehouden zal worden in 1964 te Weenen, hoopt men tot overeenstemming te komen.

STUDIEGROEP XI *(televisie)*.

De feitelijke resultaten van de vergadering van studiegroep XI behaald tijdens de Xde plenaire vergadering van het CCIR zijn geringer dan men op grond van een opsomming van de geformuleerde aanbevelingen, vragen en studieprogramma's zou vermoeden, omdat het meeste voorbereidende werk verricht werd op de CCIR-Expertsvergadering in Cannes (1961) en op de vergadering van studiegroep XI in Bad-Kreuznach (1962).

Als nieuwe onderwerpen kunnen genoemd worden de testsignalen gedurende de lijnonderdrukking van een televisiesignaal en de videostandaard voor 625-lijnen monochrome televisie voor programma-uitwisseling. Aan de Administraties is gevraagd in hoeverre hun nationale normen hieraan aangepast kunnen worden.

Wat de kleurentelevisienormen betreft, werden geen resultaten bereikt, hetgeen ook was te verwachten, aangezien de proefnemingen met NTSC- en SECAM-systemen in de EBU- en de IBTO-landen nog in volle gang zijn. Het is zelfs de vraag of op de komende speciale vergadering van studiegroep XI een beslissing genomen kan worden, maar verwacht mag worden dat tegen die tijd voldoende gegevens beschikbaar zijn om een vruchtbare gedachtenwisseling mogelijk te maken.

Op voorstel van Engeland werd besloten een speciale vergadering van studiegroep XI te houden met als onderwerp de normen voor kleurentelevisie in Europa. Deze vergadering zal in Engeland worden gehouden op data, nader vast te stellen door de voorzitter van studiegroep XI, in overleg met de betrokken Administraties, vermoedelijk in de eerste helft van 1964.

De volgende tussentijdse vergadering van studiegroep XI zal op uitnodiging van de Oostenrijkse Administratie in het voorjaar van 1965 te Wenen worden gehouden.

STUDIEGROEP XII *(omroep in de tropen)*.

Het hoofdthema was het ontwerp voor een goedkope omroepontvanger. De door de interimbijeenkomst aanbevolen ontwerpen werden hiertoe omgewerkt.

STUDIEGROEP XIII *(mobiele radiodiensten)*.

Een bijzonder actuele vraag was de invoering van selectieve oproepsystemen. Na een schifting van de ingekomen stukken is men tot de conclusie gekomen dat selectief oproepen niet alleen in de maritieme v.h.f.-band, maar ook in m.f.- en h.f.-telefoniebanden zal moeten kunnen worden toegepast, alhoewel dit laatste voorshands nog technische bezwaren zal opleveren.

Voorlopig wordt dan ook alleen aan v.h.f.-gebruik gedacht, met toepassing in de andere banden wel in de verdere toekomst.

Drie systemen zullen nader worden bekeken te weten het Amerikaanse 2-toon-systeem, het Duitse systeem met opvolgende tonen en het Japanse systeem met opvolgende dubbeltonen.

De Duitse Administratie heeft het Amerikaanse systeem onderzocht en de Amerikanen het Duitse. Aan de hand van de beschikbare gegevens is nagegaan welke kenmerkende gegevens moeten worden onderzocht en vergeleken.

Enkele landen waarvan vooral Duitsland, hebben onderzoekingen gedaan in hoeverre het mogelijk is om aan boord van schepen de richtingzoeker in de 2 MHz-band in het bijzonder op de telefonie noodfrequentie 2181 kHz, te gebruiken hetgeen in noodgevallen van veel belang is voor de zoekende schepen.

Verschillende maatregelen werden besproken om de storingen op de 500 kHz noodfrequentie zo laag mogelijk te houden.

Het Nederlandse voorstel om de uitzendingen van meteorologische kaarten, voorzover deze bestemd zijn voor schepen, te standaardiseren resulteerde in een nieuwe vraag.

In verband met het invoeren van enkelzijbandmodulatie bij de scheepvaart vraagt men zich af of een dergelijke zender met A2H respectievelijk A3H uitzending ook kan worden gebruikt voor noodverkeer m.a.w. zullen alle ontvangers en ook de automatische alarmtoestellen hierop zonder meer reageren.

De Zweden namen reeds uitgebreide proeven waarbij werd aangetoond dat dit met een normale ontvanger mogelijk is.

De ongeveer 10 verschillende dienstcodes die in de telecommunicatie worden gebruikt zijn naar aanleiding van een rapport en een recommandatie door CCITT in samenwerking met CCIR in één boek uitgegeven.

STUDIEGROEP XIV *(begripsomschrijvingen en symbolen)*.

Een belangrijke bijdrage tot de begripsomschrijvingen was het IEC-ontwerp voor een woordenboek dat aan CCIR ter kritiek werd voorgelegd. Hiertoe werd een werkgroep geformeerd die hierover zal rapporteren.

Er werd een nieuwe gecombineerde werkgroep voor tekensymbolen geformeerd uit IEC, CCITT en CCIR.

CMTT

De resultaten van de CMTT-vergadering die gelijktijdig met de 10e Plenaire vergadering van het CCIR te Genève plaats vond, kunnen in het kort als volgt worden samengevat.

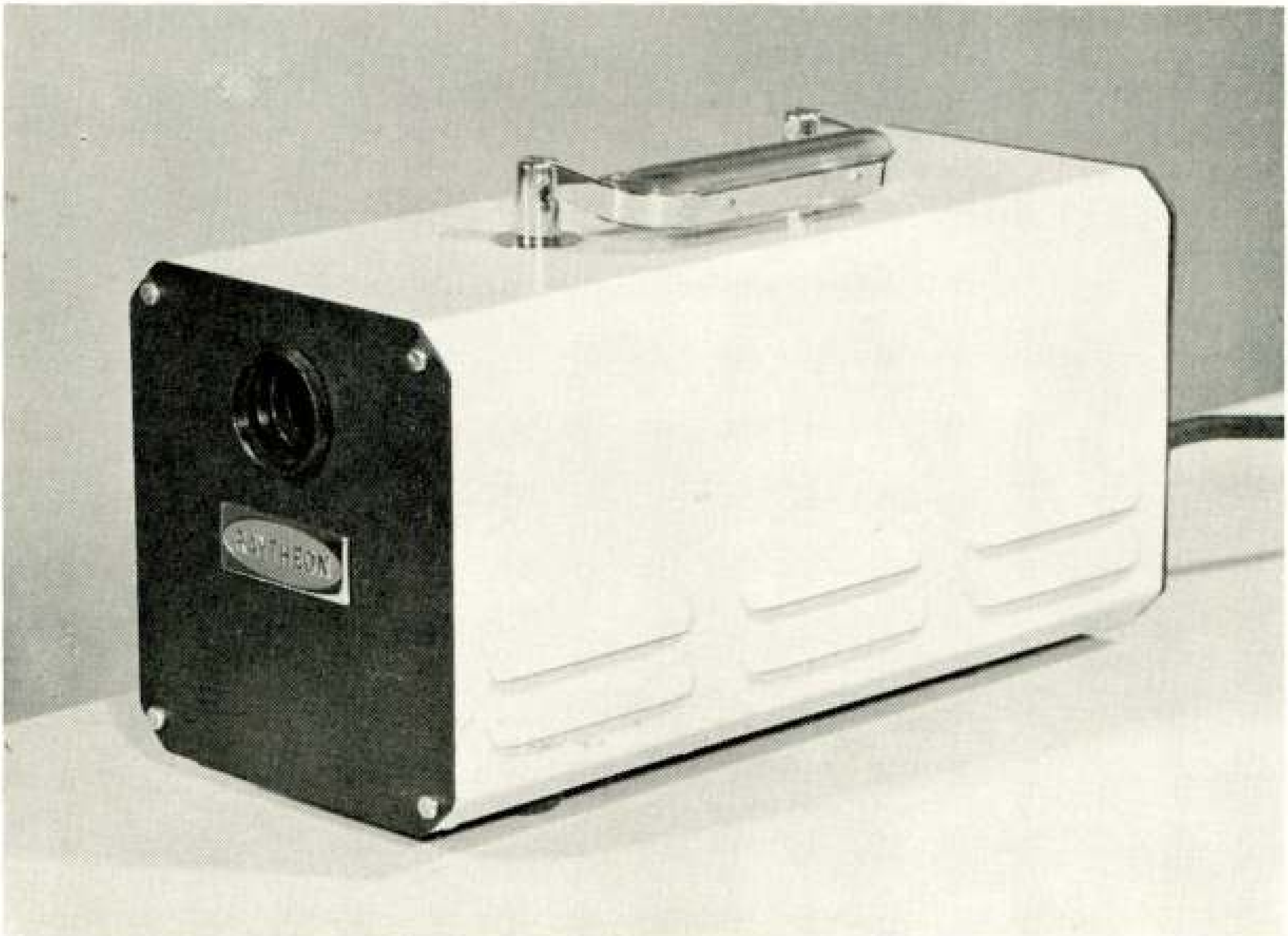
De bestaande recommandatie „Eisen voor de overdracht van monochrome televisiesignalen over grote afstanden", op enkele ondergeschikte punten gewijzigd, is aangevuld met een Annex V, waarin is aangegeven hoe de verschillende distorsieverschijnselen optellen. Hiermede wordt vastgelegd, welke eisen gesteld moeten worden voor trajecten, die korter of langer zijn dan het referentiecircuit van 2500 km.

In de vorm van een rapport is een aanduiding gegeven in hoeverre de eisen voor overdracht van kleuren-televisiesignalen afwijken van die voor monochrome signalen.

Voorts is aanvaard de vraag om na te gaan, of de huidige definitie van het referentiecircuit (2500 km met twee tussenliggende videopunten) passend is om voldoende resultaten te waarborgen op lange, werkelijke circuits, waar eventueel meer videopunten aanwezig kunnen zijn.

Een nieuw studieprogramma is aangenomen ter bestudering van de mogelijkheid om fundamentele kwalitatieve parameters van een televisieverbinding automatisch op afstand te controleren, door breedbandige meetsignalen om te zetten in smalbandige informatie.

GAS-LASER VAN RAYTHEON



Raytheon introduceerde een nieuwe gas-laser werkend op een golflengte van 3,51 μ. Deze in het midden van het infrarood gelegen golflengte correspondeert met een minimum-absorptie-venster in de atmosfeer en is daarom ook geschikt voor communicatietoepassingen. Voor dit doel kan een modulator worden ingebouwd.

De 6 inch kwartsbuis in de laser heeft een gasvulling van helium-xenon en is voorzien van Brewster-vensters. Het uitgangsvermogen bedraagt 0,25 mW terwijl de divergentie van de vlakke bundel zonder collimatie 20 boogminuten bedraagt. De bandbreedte bedraagt 110 MHz.

Het apparaat is voorzien van ingebouwde lichtnetvoeding en weegt ongeveer 4,5 kg.

CONGRESSEN E.D.

Internationaal symposium PTGMTT 1964.

In New York wordt van 19-21 mei 1964 het jaarlijkse symposium van de Professional Group on Microwave Theory and Technics van het IEEE gehouden. Bijzondere aandacht wordt ditmaal gevraagd voor lasers, groot-vermogen-technieken, millimeter- en submillimeter-golftechnieken en componenten, vaste-stof-toepassingen.

Voorzitter van de commissie voor het technisch programma: Leonard Severn, 1964 PTGMTT International Symposium, Sperry Gyroscope Company, Great Neck, L.I., N.Y.

BOEKAANKONDIGINGEN ENZ.

In de serie Technische Mitteilungen Halbleiter van Siemens & Halske A.G. verschenen zes nieuwe uitgaven n.l. „Probleme der Zuverlässigkeit von Halbleiter-Bauteilen", „U.H.F.-Tuner mit Mesa-Transistor AF 139", Meszverfahren für

Groszsignal-Kenngröszen von Transistoren, Wärmeableitung bei Transistoren, Paarung von Transistoren für NF-Gegentaktendstufen en A-Endstufe mit verbessertem Leistungstransistor.

BOEKBESPREKING

Servicing transistor radios and printed circuits by Leonard Lane (edited by E. A. W. Spreadbury), Iliffe Books Ltd, Londen, 1962. Afmetingen 8¾" x 5½", 264 blz. Prijs 42 sh.

Dit boek is geschreven om de doorsnee service-technicus de weg te wijzen in de problemen, die de introductie van transistors en gedrukte bedrading voor hem meebrengen. Wiskundige formuleringen zijn zorgvuldig vermeden, wat veel mensen van de praktijk een grote verdienste zullen vinden.

In de eerste zes hoofdstukken (124 blz.) wordt een inzicht gegeven in de werking van transistors en hun toepassing in radioapparaten. Alles wat te pas komt in een moderne superheterodyne ontvanger passeert daarbij de revue.

De hoofdstukken 6 t/m 10 geven dan een gedetailleerd overzicht van de methoden om eventuele fouten op te sporen en te verhelpen. De schrijver richt zich tot technici die met buizen-apparaten vertrouwd zijn. Daardoor wellicht gaat hij minder uitvoerig in op specifieke stoorzoeksystemen, zoals „signal tracing" en „signal injection", al komen beide impliciet wel aan de orde.

Bij de autoradio worden ook de storingen behandeld die het ontstekingsmechanisme kan veroorzaken, alsmede de middelen om deze storingen te onderdrukken.

Het elfde hoofdstuk is geheel gewijd aan de gedrukte bedrading en geeft aanwijzingen hoe te handelen wanneer aan de bedrading zelf, of aan de erin opgenomen componenten iets defect is.

Hoofdstuk 12 geeft een overzicht van de verschillende transistors.

Een appendix van 7 bladzijden vat nog eens alles samen, wat bij het opsporen van fouten in radioschakelingen nuttig kan zijn.

De Engelse bewerker heeft de woordkeus van het oorspronkelijk Amerikaanse boek aangepast aan Engels gebruik en voor foto's van onderdelen uit de Engelse industrie gezorgd.

De technicus die zich met radioreparatie bezig houdt zal in dit boek veel handige wenken vinden.

D. H. J.

Uit het N.E.R.G.

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