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Effects of configuration parameters on the radiation of the ISDN S-bus

Norp, A.H.J.

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EFFECTS OF CONFIGURATION PARAMETERS
ON THE RADIATION OF THE ISDN S-BUS.

by
A.H.J. Norp.

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Professor : prof. ir. J. de Stigter.

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1. INTRODUCTION.

In the nearby future, ISDN services will be introduced in the Dutch telephone network. One of the main requirements for ISDN services is a fully digital network. For the in-house part of this digital network, the international standardization organization CCITT has defined the S-bus. The S-bus is a digital bus system, to which, next to the usual telephone set, a range of terminals (e.g., fax, telex, and computers) can be connected. To make the introduction of ISDN economically feasible, the S-bus is designed to be operable on existing voice grade telephone cables. The bandwidth of a voice channel in the existing telephone network is approximately 4 kHz; the 192 kbit/s digital signals on the S-bus require a much larger bandwidth. Because of the relative high bandwidth and bit rate, attention should be paid to the EMC (Electro-Magnetic Compatibility) and EMI (Electro-Magnetic Interference) aspects of the S-bus.

Recent studies on the EMI aspects of the S-bus [5,6,7,8,26] have indicated that S-bus signals may cause interference problems. In worst case situations within the CCITT recommendations, radiation levels will exceed the limitations imposed by national EMC regulations [26]. Especially the interference to long and medium wave broadcast may be significant. Not only the interference to other equipment is important: emitted signals could also be received with a small antenna, and it might be possible to recover the original transmitted information. In experiments, Daneffel [5] was able to receive signals that were easily recognizable with only a simple antenna.

Because of the sensitive nature of the information transmitted, COMSEC (COMmunication SECurity) aspects of the S-bus are very important. An ordinary telephone conversation will not be interesting to third parties, but eavesdropping on a contracting firm could be highly profitable to the competition. Other types of
sensitive information are PIN-codes in electronic banking and passwords in electronic data traffic.

Because radiation of the S-bus proves to be a problem, it is interesting to know which parameters of the S-bus configuration determine the radiation. Which parameters are important to specify a S-bus with a low radiation level and which parameters can be disregarded. Even within the CCITT recommendations, the design of an actual S-bus implementation leaves open a number of choices that influence the radiation. The purpose of this thesis is to theoretically investigate the influence of various configuration parameters. The main goal is not a quantification of the level of radiation, but qualitative insight in the effects of the parameters.
2. THE ISDN S-BUS.

2.1. Introduction.

In the definition of ISDN in the CCITT recommendations, a whole new digital user-network interface is specified. In Recommendation 1.411, a reference configuration of this user-network interface is defined. As shown in figure 2.1., the reference configuration is divided into three functional entities. Between these elements three reference points are defined. The U-reference describes the two conductor transmission line between the local exchange and the entry of the user premises. This subscriber line is connected to a NT1 (network termination), which adapts the signals on the subscriber line to the signals on the in-house installation. Via the T-reference, the NT1 is connected to the NT2. The NT2 distributes the signals from the NT1 via the S-reference over a number of ISDN terminals (TEs). In a simple in-house installation, the NT2 is formed by a bus system. This bus system is called the S-bus. In more complex in-house installations, a private exchange (PABX) will be used for the distribution. From the PABX onwards, a S-bus configuration is possible again.

![Figure 2.1: Reference configuration for user-network interfaces.](image)

The S-bus provides each terminal connected, with a basic access to the network. The basic access, as defined in Recommendation 1.412, consists of two B-channels of 64 kbit/s and a D-channel of 16 kbit/s in both transmission directions. The B-channels are used for the transfer of user information; the D-channel is mainly used for signaling, but may also transfer user information.
One of the mayor features of the S-bus is its flexibility. It can transfer information to and from a range of terminals, varying from simple telephone sets eg. to much more complex fax equipment. Because the S-reference is implemented in every terminal that is connected to the bus, the S-bus system has to be of limited complexity and cost. A mayor increase in cost of especially the telephone set, which is the basic ISDN terminal for almost every user, would be unacceptable. In the specification of the ISDN recommendations, this restriction has been taken in consideration.

In the context of this thesis, which deals with EMC aspects of the S-bus, the Layer-1 characteristics of the S-bus are of most interest. These characteristics are specified in the CCITT Recommendation I.430. In the remaining paragraphs of this chapter, attention will be focused on these Layer-1 characteristics.

2.2. Wiring configurations.

In principle, terminals can be connected everywhere on the bus. There are, however, some restrictions to the maximum length of the bus and the maximum number of terminals that can be connected. The restrictive factors are the maximal differential round trip delay and attenuation. The differential round trip delay is limited to a half bit interval, in order to maintain synchronization between transmissions coming from different TEs. Considering the differential round trip delay, the short passive bus has been defined. This bus configuration (depicted in figure 2.2.(a) ) is maximally 300 meters long. The maximum number of TEs that can be connected is eight. The bus length can be extended when the TEs are grouped together. Figure 2.2.(b) shows the so called extended passive bus configuration, in which the maximum distance between any two TEs is limited to 50 meters. The maximum distance between the first TE and the NT can thus be extended to 500 meters. When only one TE is connected, as in the point-to-point configuration, there is no differential round trip delay. Then the maximum attenuation
becomes the restrictive factor. The maximum attenuation is 6 dB, which allows for a bus length of 1 kilometer (figure 2.2.(c)).

The S-bus (point-to-point configuration) is shown in more detail in figure 2.3. It consists of two cable pairs, one for each transmission direction (Space-Division-Multiplex). Both cable pairs are balanced with the use of two balance-unbalance transformers and differential mode terminated on each side in a 100 Ω resistor. This 100 Ω resistor represents an approximately matched termination.
The use of balance-unbalance transformers offers the possibility to provide power to the terminals via a phantom circuit. In the phantom circuit, a power source is connected between the center taps of the transformers on the NT side. Power is thus available between the center taps on the TE side.

2.3. Functional characteristics.

The digital signals on the S-bus are divided into frames: for each transmission direction in a different frame structure. These two frame structures are shown in figure 2.4. Each frame contains a total of 48 bits in a frame interval of 250 $\mu$s, which gives a gross bit rate of 192 kbits/s. The net bit rate of a basic access is $(2 \times B + D)$ 144 kbits/s. The frames are transmitted using Alternate Mark Inversion (AMI) line coding. In the AMI strategy, a binary zero is represented by an alternating positive or negative pulse. No pulse is transmitted for a binary one. A violation of this code rule is used to indicate the start of a frame for frame alignment. When an even number of zeros is transmitted, no DC component is present in the signal. An even number of zeros is always provided by the balancing bits that are inserted in the frame structure. In the direction from NT to TE, every frame is balanced. In the direction from TE to NT, each channel is balanced separately, because separate channels can be used by different TEs.
2.4. Electrical characteristics.

The output pulse shape for the ISDN S-bus is specified in CCITT Recommendation I.430. Except of overshoot, output pulses should be within the mask shown in figure 2.5. Overshoot is permitted up to
5% of the amplitude at the middle of the pulse, provided that the overshoot has a duration less then 0.25 μs. The nominal pulse amplitude is 750 mV, zero to peak. A clean square wave with nominal amplitude and a pulse width of 5.21 μs will best fit the mask. A very short rise time, however, causes very high frequency contents, which is disadvantageous from an EMC point of view. The slowest possible rise time which will fit the mask is approximately 1.0 μs.

![Diagram of Transmitter output pulse mask.](image)

*figure 2.5:* Transmitter output pulse mask.

The receiver input impedances of the NT and TE are specified by two slightly different impedance templates. The impedances are required to exceed the impedances indicated by these templates. The impedance of both templates is 2500 Ω over a specified frequency range; decreasing 20 dB/dec on both sides of this frequency range. The impedance templates only specify impedances in the range of 2 kHz to 1 MHz. Outside this range no impedances are given. The same templates are used to specify the output impedances of the transmitters. When transmitting a binary one (no pulse), the transmitters are required to have a output impedance exceeding the templates. When transmitting a binary zero, a minimum output impedance of 20 Ω is specified.
The nominal voltage of the power source in the phantom circuit is defined by the CCITT recommendations as 40 Volts. The power consumption of a TE is limited to 1 Watt. Parameters of the power source and sink that affect the transmission characteristics of the bus, such as input and output impedances at signal frequencies, are not specified in the CCITT recommendations. These parameters, however, can influence the level of radiation and should therefore be accounted for.

The level of radiation is for a great deal determined by the unbalance properties of the S-bus. These properties are specified in the CCITT recommendations by two unbalance parameters: the Longitudinal Conversion Loss (LCL) and the output signal balance. LCL is a measure of mode conversion; when e.g. a poorly balanced termination is excited by a common-mode signal, this signal will be partly converted to a differential-mode voltage. The LCL is the ratio of the exciting common-mode voltage and the developed differential-mode voltage, expressed in dB's. For the input and output ports of the NT and the TE a minimum LCL of 54 dB is required in the frequency range from 10 kHz to 300 kHz; decreasing 20 dB/dec for higher frequencies up to 1 MHz. For frequencies above 1 MHz no minimum LCL is given. For interconnecting media (the cable) a minimum LCL of 43 dB is specified at a spot frequency of 96 kHz.

The output signal balance is a measure for the amount of unwanted common-mode signal generated by a differential-mode source. It is defined as the ratio of the differential-mode voltage and the common-mode voltage, expressed in dB's. For frequencies up to 96 kHz the minimum output signal balance of the NT and TE output ports is 54 dB. For frequencies in the range from 96 kHz to 1 MHz a decrease of 20 dB/dec is permitted. The output signal balance is not specified for frequencies above 1 MHz.
The CCITT recommendation I.430 lacks a specification of the type of cable that is to be used. Since the ISDN S-bus was designed to operate on existing subscriber premises cables, the use of conventional unshielded telephone cables can be assumed. The type of cable may vary according to national regulations, but usually a PVC insulated twisted quad cable is used. This type of cable is, due to the cable twist, fairly well balanced. When installed in actual buildings, the balance of a twisted cable may be impaired by metal structures such as other cables, wiring, or plumbing. Recent investigations [17] have indicated that typical in situ quad cables have LCL values of approximately 50 dB. At 96 kHz LCL values more than 25 dB better than the value of 43 dB specified by the CCITT were found.
3. MODELLING THE S-BUS.

3.1. Introduction.

Although the main part of the ISDN S-bus specifications are laid down in CCITT recommendation I.430, the design of the actual implementation leaves open a number of choices which can greatly influence the electromagnetic interference of the S-bus. Experience has shown that modification of various parameters, such as source and load impedances and wire length, can have a significant effect on the level of radiation of a transmission system. A theoretical estimation of these effects can be very useful in selecting a proper set of parameters to minimize radiation of the S-bus. To get a theoretical estimation of the radiation, a model of the S-bus is needed. By varying different parameters of the transmission system in this model, a qualitative insight in the influence of these parameters on the level of radiation can be obtained. A quantification of the radiation is useful, but not the main goal of the model.

At frequencies of interest in the ISDN S-bus, radiation is predominantly caused by currents. The magnitude of these currents, together with the loop they describe, determines the level of radiation. A distinction should be made between common-mode and differential-mode radiation, caused by the common-mode and differential-mode current loop respectively. Ideally, the S-bus is perfectly balanced and carries no common-mode currents. However, some residual unbalance of the NT and TE equipment is tolerated by the S-bus specifications. Also can the cable exhibit some unbalance due to imperfect cable manufacturing and parasitic capacitance to other cables and metal structures. Differential-mode currents are always present, since the information signal is transmitted in this mode. Because the differential-mode current loop encloses only a small area, the differential-mode radiation is usual not very serious. Moreover is the differential-mode
radiation reduced by the twisting of the cable. The common-mode current encloses a much larger area, because it returns through a ground conductor and not through the cable. So, even if the common-mode current is much smaller than the differential-mode current, the common-mode radiation will be the principle source of interference.

Since most attention should be paid to the common-mode radiation, the model should accommodate the non-ideal properties of the S-bus that cause the common-mode current. Unbalances of the cable, source and load impedances, line drivers, etc. will have to be modelled. Also a ground conductor, through which the common-mode current returns, should be adopted in the model.

The ISDN S-bus signal contains frequencies up to a few Megahertz, and due to fast clock signals in the TE and NT equipment even higher frequencies can be present. A frequency of 1 MHz corresponds to a half wavelength of 150 m, which is already smaller than the maximum length of a short passive bus. So, especially at higher frequencies, resonances can occur. The model that is used has to be suitable to predict these resonances. Cross-sectional dimensions of the bus can be assumed much smaller than the wavelength.

3.2. Modelling of the S-bus configuration.

Before an appropriate model can be chosen, a closer look has to be taken upon the configuration that has to be modelled. The basic S-bus configuration is a point-to-point configuration, with only one TE and one NT. Usually there are more TEs connected to the bus. But, since the addition of more terminals on various points of the bus would make the model more complicated, only the basic configuration is modelled. The point-to-point cable configuration is depicted in figure 2.3.
Because a quantification of the radiation is not considered an important goal in the modelling of the S-bus, the four-wire configuration is simplified to a two-wire model. The exact level of radiation of the S-bus can no longer be calculated with a two-wire model, but a better insight can be obtained in mechanisms that determine the level of radiation. On the four-wire bus, information is transported in two directions; now only one transmission direction remains. Therefore, a more general transmission line is modelled.

A model of the two-wire bus can be obtained simply by removing one pair out of the four-wire configuration in figure 2.3. The only problem that is encountered is the adaptation of the phantom power circuit. The power source and sink in the phantom circuit are connected between the two wire pairs. No galvanic contact is present between the bus and ground: the bus is entirely floating. Substituting the power source and sink by a source and sink between the remaining wire pair and the ground reference is therefore not correct. At signal frequencies, however, the power source and sink can be represented by an impedance to the ground reference because of existing capacitances to earth. The thus created two-wire model is depicted in figure 3.1. It is important to notice that the ground reference on the receiver side is not the same as on the transmitter side. The two grounds are interconnected by a ground conductor, which greatly influences the level of radiation.

![Diagram of two-wire model](attachment:two-wire_model.png)

*figure 3.1: Two-wire model.*
The unbalance of the NT and TE ports, described by the output signal balance and the Longitudinal Conversion Loss (LCL), cause a great deal of the radiation of the S-bus. The transmitters in the TE and NT generate a differential-mode signal. But, because of an imperfect output signal balance, a common-mode component is also transmitted. The LCL is normally used as a measure for the differential-mode voltage that is developed in a port when it is excited by a common-mode signal. By reciprocity, it can also be used as a measure of the common-mode voltage that is generated in a TE or NT port excited by the differential-mode S-bus signal. The S-bus signal is, when it is reflected at a TE or NT port, partly converted into a common-mode signal. The common-mode component of the reflection gives rise to an increase in radiation.

The unbalance of the NT and TE ports is caused by asymmetry. The centre tap of a transformer may not be precisely in the middle of the coil, or the termination of two wires may slightly vary due to different capacitances, etc. This asymmetry has to be brought into the model, preferably in such a way that the LCL and output signal balance can be easily modified.

Another important property of the NT and TE ports, in regard to radiation, is the input or output impedance. Together with the terminating resistor it forms the termination of the transmission line and is therefore determinant for the reflections on that line. The differential-mode impedance (i.e. the impedance seen between the wires) is easily recognisable: it is formed by the terminating resistor paralleled with the input or output impedances of the receiver or transmitter. These impedances are specified by the CCITT. The common-mode impedance (i.e. the impedance seen between the wires and ground) is more difficult to determine. Looking at figure 3.1., it can be seen that the impedance to ground representing the power source and sink is part of the common-mode impedance. However, because of capacitances between the primary and secondary coils of the transformer, an
impedance to ground can also be seen through the transformer. The values of both impedances comprising the common-mode impedance are not specified by the CCITT recommendations and are highly dependent of the actual implementation of the TE or NT.

A convenient way to model the unbalance and terminating impedance of the NT and TE ports is shown in figure 3.2. The receiver is modelled by an equivalent circuit composed of three impedances in a Π-circuit. Every passive network that is terminating a two-wire line (plus ground conductor) can be represented by such an equivalent circuit. For a receiver this equivalent network suffices; for a transmitter it does not. A transmitter also serves as a line driver, while a receiver only terminates the line. Therefore active components have to be added to the equivalent circuit of a transmitter. Because the unbalance of the transmitter has to be taken into account, two voltage sources are needed. With the equivalent circuit in figure 3.2. representing the transmitter, all five independent variables characterizing the transmitter can be modelled.

\[ \begin{align*}
Z_{01} & \quad Z_{0m} \quad Z_{1m} \quad Z_{12} \\
V_1 & \quad V_2
\end{align*} \]

figure 3.2.: Two wire model with termination represented by an equivalent circuit.

Now the NT and TE ports are represented in the model, attention has to be directed to the modelling of the interconnection media. The NT and TE ports are interconnected not only by the cable, but also by the ground conductor. In a real situation the ground conductor is not always easily recognized. It can be formed by the ground lead of the power supply wiring when the NT and TE both
have an external power supply. But a TE that derives its power from the phantom circuit might not have any other connection than the S-bus cable. Even in this case a ground conductor (or better a ground path) is present. The TE is capacitively coupled to nearby metal structures such as the metal reinforcement in a concrete floor. The NT can also be capacitively coupled to this metal structure, and thus a ground path can be formed.

If there is no real metal conductor that connects the NT and the TE, it is difficult to localize the common-mode current loop. It is, however, important to know the size of the loop area, since the common-mode radiation depends on it. Therefore a real S-bus configuration is very difficult to model if no clear ground conductor is present. In fact, only simple ground conductor configurations are sufficiently easy to model, so usually a simplification has to be made. In the model that is used here, the ground conductor is represented by an infinite, perfectly conducting metal plate. The cable is running parallel to this plate. The impedance of the ground path, which in reality is not perfectly conducting, is assumed to be concentrated at the NT and TE.

The last item remaining to be modelled is the cable. In the S-bus typically an unshielded, twisted cable is used. An ideally twisted cable running parallel to a perfectly conducting metal plate is perfectly balanced. A normal twisted cable, however, is not
perfectly balanced, because it is surrounded by all kinds of metal structures and is not perfectly twisted. If unbalance of the cable has to be taken into account, a twisted cable cannot be used in the model. Therefore, the twisted wire configuration is here substituted by a parallel wire configuration as shown in figure 3.3. With a parallel wire, unbalance is created if one of the conductors is closer to the metal plate than the other. The use of parallel wires instead of twisted wires not only increases the common-mode radiation, but also causes more differential-mode radiation. Because this type of radiation can normally be neglected, it is also neglected in the model.

3.3. Selection of a suitable method for calculating the radiation.

In the previous paragraph, a two-wire model has been constructed of the S-bus configuration. Now a method of calculation has to be found to determine the level radiation of this model configuration. In literature, a number of methods is described for calculating the radiation of wires above a ground plane. Three of these methods will be briefly discussed here, and a comparison of the suitability will be made.

The simplest model used for calculating the radiation is the short-wire model [21,32]. In this model, the wire length is assumed short compared to the wavelength. Transmission line parameters, such as propagation time, reflection, and resonance, are neglected. The assumption is made that the current distribution is constant at all points along the transmission line. The ISDN S-bus can not be considered electrically short, and therefore the short-wire model is not suitable for calculating the radiation of the S-bus.

A calculating method that does account for transmission line parameters is the Transmission Line Modelling method (TLM). TLM is a time domain method for the analysis of transmission lines. It
was originally defined for two conductor lines but, using matrix calculation, the method can also be applied to a multiple conductor line [1,10,18].

With the TLM method, the level of radiation is calculated in two steps. In the first step, the current distribution along the transmission line is calculated. Here the assumption is made that the propagating wave has an essentially TEM (Transversal Electro-Magnetic) character. This is only true for frequencies that are not too high. The transmission line and the electromagnetic properties of the surrounding media are modelled by infinitely small transmission line sections. Every transmission line section is then represented by a local equivalent circuit. In the second step, the contribution to the radiation of the local current in each transmission line section is determined. With an appropriate time retarded integral, the total radiation is then calculated.

The TLM method is not entirely consistent: the radiation is computed with use of the calculated current distribution, while this distribution is calculated neglecting the radiation. However, usually only a relative small portion of energy is radiated, and the TLM method provides fairly good results.

The third calculation method is the Method of Moments (MOM) technique [2]. With this method, the problem is modelled in a virtually exact fashion: the various assumptions inherent in the TLM method are not employed. Just like in the TLM method, the radiation is determined in two steps. In the first step, the current distribution is calculated. In the second step, the resulting radiation is computed. The calculation of the radiation can be done in the same way as in the TLM method. The difference between the MOM method and the TLM method lies in the first step. In the TLM method, the transmission line is divided into sections; in the MOM method, the current distribution is divided. The current distribution is expanded as a linear combination of N known basis distribution functions. The problem is thus reduced to
the calculating of \( N \) unknown constants. This leads to one equation with \( N \) unknowns. To resolve the \( N \) constants, \( N \) linearly independent equations are needed. For this, a set of weighting functions is used. These functions represent the electromagnetic boundary conditions at various points of the transmission line. For both the basic functions as the weighting function, different functions can be chosen. The accuracy of the MOM method is greatly influenced by the choice of basic and weighting functions.

Comparing the three methods, one can easily see that the short-wire model is not suitable: the restriction of an electrically short wire can not be met. Choosing the most suitable method out of the remaining two is somewhat more difficult. The MOM method seems better, because it is not restricted to quasi-TEM waves like the TLM method. However, the MOM method has some major setbacks. The MOM technique uses large matrix calculations and is therefore not very efficient. The TLM method has a much higher computational efficiency. Moreover, the MOM method is not suitable when radiation has to be predicted over a large frequency range (research has indicated that the MOM method suffers from numerical instabilities at low frequencies [22,30]). The TLM method, on the other hand, can be used as long as the cross sectional dimensions of the transmission line can be assumed small compared to the wavelength. In the S-bus situation, with typical cross sectional dimension of one meter, it can be used way above frequencies of concern. Furthermore, the TLM method provides insight in the mechanisms that determine the radiation, while with the MOM methods only numerical results can be calculated. Summarizing, the TLM method is more useful, especially in the lower frequency range. This method is therefore selected for the calculation of the current distribution.

3.4.1. Introduction.

In this paragraph, the TLM theory of wave propagation on lossless multiconductor transmission lines is presented. This theory is based on the TLM theory for two-conductor transmission lines. To accommodate more conductors, matrix calculations are used. Because of this matrix formalism, the same notation of relationships between voltage and current waves and their reflections can be used as for the two-conductor line. A major difference between the multiconductor and the two-conductor theory is the introduction of propagation modes.

The transmission line considered here is a lossless line consisting of \( n+1 \) conductors. One of these conductors is chosen as the ground reference. The transmission line is assumed uniform along its length. The cross-sectional configuration, however, can be arbitrary. In particular, the dielectric material may be inhomogeneous.

The use of the TLM theory is restricted to the propagation of TEM waves. A wave propagating on a transmission line with homogeneous dielectric material is a TEM wave. But if the dielectric material is inhomogeneous (e.g. when the conductors are separately isolated), the propagation is TM, not TEM. In the low-frequency range, this TM wave looks very much like the TEM wave: it is "quasi-TEM". Therefore, though this is not entirely correct, the TLM theory is also used on transmission lines with inhomogeneous dielectric materials.

3.4.2. Derivation of the propagation modes.

The basis of the TLM theory is found in the telegrapher's equations. These equations are:
\[
\frac{\partial V(z,t)}{\partial z} = -L \frac{\partial I(z,t)}{\partial t} \\
\frac{\partial I(z,t)}{\partial z} = -C \frac{\partial V(z,t)}{\partial t}
\]  

Here \( V \) and \( I \) are \( n \)-dimension column vectors

\[
V = \begin{bmatrix} V_1 \\ V_2 \\ \vdots \\ V_n \end{bmatrix} \quad I = \begin{bmatrix} I_1 \\ I_2 \\ \vdots \\ I_n \end{bmatrix}
\]

with \( V_i \) and \( I_i \) the voltage and current on the \( i \)th conductor. \( L \) and \( C \) are \( n \times n \) matrices

\[
L = \begin{bmatrix} L_{11} & L_{12} & \cdots & L_{1n} \\ L_{21} & L_{22} & \cdots & L_{2n} \\ \vdots & \vdots & \ddots & \vdots \\ L_{n1} & L_{n2} & \cdots & L_{nn} \end{bmatrix} \quad C = \begin{bmatrix} \sum_{i=1}^{n} C_{1i} & -C_{12} & \cdots & -C_{1n} \\ -C_{21} & \sum_{i=1}^{n} C_{2i} & \cdots & -C_{2n} \\ \vdots & \vdots & \ddots & \vdots \\ -C_{n1} & -C_{n2} & \cdots & \sum_{i=1}^{n} C_{ni} \end{bmatrix}
\]

with \( L_{ij} \) the mutual inductance between conductor \( i \) and conductor \( j \), and \( L_{ii} \) the self inductance of conductor \( i \), both per unit of length. \( C_{ij} \) is the capacitance between conductor \( i \) and conductor \( j \), and \( C_{ii} \) is the capacitance between conductor \( i \) and ground, also per unit of length.

A propagation mode is a solution to the telegrapher's equations (1) and (2), with current and voltage vectors in the following form:

\[
V(z,t) = V \cdot f(z - ut) \\
I(z,t) = I \cdot f(z - ut)
\]

Here \( V \) and \( I \) are constant vectors. Substitution of these current
and voltage vectors in the telegrapher's equations (1) and (2) results in the following relations:

\[ V = \varepsilon LI \]  
(3)

\[ I = \varepsilon CV \]  
(4)

From equation (3) and (4) the current vector \( I \) can be eliminated, resulting in an eigenvalue equation for \( V \)

\[ (LC)\varepsilon^2 V = \frac{1}{\varepsilon^2} V \]  
(5)

where \( 1/\varepsilon^2 \) is an eigenvalue of the matrix \( LC \), and \( V \) is the associated eigenvector. When \( V \), instead of \( I \), is eliminated from (3) and (4) an eigenvalue equation for \( I \) results. A set of eigenvectors \( I \) can then be found, related to the eigenvectors \( V \) through equation (4). Both \( I \) and \( V \) correspond to the same eigenvalues \( 1/\varepsilon^2 \).

In general, there are \( n \) distinct eigenvalues. Each of these eigenvalues corresponds with a propagation mode. In case the dielectric is homogeneous, all propagation modes coincide, and all eigenvalues are identical. This occurs when \( LC=(1/\varepsilon^2)U \), with \( \varepsilon=1/\sqrt{\mu\epsilon} \) is the propagation velocity, and \( U \) is the identity matrix. Any vector \( V \) will then satisfy equation (5).

3.4.3. The characteristic impedance and admittance matrices.

In homogeneous dielectrics, \( \varepsilon \) is a common propagation velocity for all modes. Therefore, for waves travelling in positive direction, equations (3) and (4) can be written as

\[ I = \varepsilon CV = \frac{1}{\varepsilon} L^{-1} V = Y_0 V \]

where \( Y_0 \) is the characteristic admittance matrix. In inhomogeneous dielectrics, there is no common velocity for all modes, and (3)
and (4) are only valid for one propagation mode. However, equation (3) and (4) can be written as

\[ M_1 = CM_v [u] = L^{-1} M_v \begin{bmatrix} 1 \\ \frac{1}{u} \end{bmatrix} \]  

(6)

Here \( M_v \) and \( M_1 \) are matrices whose columns are the voltage and current eigenvectors, \([u]\) and \([1/u]\) are matrices with the (inverse of the) propagation velocity \( u \) of each mode on the diagonal.

The characteristic admittance matrix for a line with inhomogeneous dielectrics is defined by

\[ M_1 = Y_0 M_v \]  

(7)

From equations (6) and (7), the following expression for \( Y_0 \) can be found:

\[ Y_0 = CM_v [u] M_v^{-1} = L^{-1} M_v \begin{bmatrix} 1 \\ \frac{1}{u} \end{bmatrix} M_v^{-1} \]  

(8)

The characteristic impedance matrix is found simply by inversion of the admittance matrix

\[ Z_0 = Y_0^{-1} \]

Because any unidirectional wave can be expressed as a linear combination of propagation modes, the following relation between the current and voltage in a forward wave can be found:

\[ I_f = Y_0 V_f \]  

(9)

or

\[ V_f = Z_0 I_f \]  

(10)

Similarly for a backward wave:

\[ I_b = -Y_0 V_b \]  

(11)
or

\[ \mathbf{V}_b = -Z \mathbf{I}_b \]  \hspace{1cm} (12)

The admittance matrix \( \mathbf{Y} \) and the impedance matrix \( Z \) describe the matched termination network. If a transmission line can be terminated with a matched termination network, reflections can be prevented.

3.4.4 The equivalent circuit of a multi conductor transmission line.

The matched termination network is also used to represent a transmission line by an equivalent circuit. A matched termination is, by definition, equivalent to the transmission line extending to infinity [1]. However, a transmission line is always of a limited length, and reflections can be present on the transmission line. These reflections cannot be represented by the matched termination network alone. If reflections are present (and to calculate reflections), active components have to be added to get the equivalent circuit that shown in figure 3.4.(b).

\[ \text{figure 3.4.} \text{: Multiple conductor transmission line and its equivalent circuit: (a) transmission line terminated into an arbitrary network, (b) equivalent circuit.} \]
Figure 3.4.(a) shows a transmission line with an incident wave \( V, I \) terminated into an arbitrary network. The incident wave is partly reflected at the arbitrary network. The sum of the incident wave and its reflection determines the voltage and current vectors at the end of the line \((z=l)\).

\[
V_l = V_{fl} + V_{bl} \\
I_l = I_{fl} + I_{bl}
\]

(13) (14)

The backward wave can be eliminated using equation (10) and (12). The resulting equation is:

\[
V_l + Z I_l = 2V_{fl}
\]

(15)

which is exactly the equation that describes the response of the equivalent circuit in figure 3.4.(b). For a backward incident wave at the begin of the transmission line, the forward wave can be eliminated, resulting in:

\[
V_o - Z I_o = 2V_{bo}
\]

Thus, the transmission line is represented by two equivalent circuits, one for each end of the line. These equivalent circuits consist of a matched termination network, with \( n \) voltage sources \( 2V_{bo}, 2V_{fl} \) connected in series at the terminals. The voltage sources \( 2V_{bo} \) at the beginning of the transmission line represent an incident backward wave. But when a wave is first transmitted, there is no incident backward wave because the wave is not yet reflected at the end of the line. Therefore, the starting forward wave \( V_{fo} \) is calculated with only the voltage source of the transmitter active.

Knowledge of \( V_o \) and \( I_o \) at the begin of the transmission line and of the characteristic impedance matrix of the line is sufficient to determine the forward wave \( V_{fo} \). This wave propagates towards the end of the line. In general, \( V_{fo} \) is a combination of different propagation modes. Because these modes can have different propaga-
tion velocities, they arrive with different delays at the end of the line. Due to this mode dispersion, knowledge of \( V_{fo} \) alone is not sufficient to determine \( V_f \) along the line. The wave \( V_{fo} \) has to be decomposed into propagation modes. Then the transit time for each mode has to be determined. The forward wave at any point along the line can be obtained by summing the propagation modes at their appropriate transit time.

3.4.5. The reflection and transmission matrices.

When the forward wave arrives at the termination of a transmission line, this \( V_f \) is partly reflected. This reflection can be described using reflection and transmission matrices, analogous to the reflection and transmission coefficients for a two-conductor line.

Consider a transmission line with a termination that is described by the impedance matrix \( Z \). The current and voltage vectors at the end of the line must then satisfy

\[
V = Z I
\]

This equation is used to get

\[
V = 2(U + Z Z^{-1})^{-1} V_f
\]  \hspace{1cm} (16)

from equation (15). Equation (16) has the desired form

\[
V = \tau V_f
\]  \hspace{1cm} (17)

with

\[
\tau = 2Z(Z + Z^{-1})^{-1}
\]

the transmission matrix.
The reflection matrix $\rho_n$ is defined by

$$V_{bl} = \rho_n V_{fl}$$

(18)

From (13) and (17), the reflected signal can be written as

$$V_{bl} = V_i - V_{fl}$$

$$= (\tau_n - U) V_{fl}$$

So, the reflection matrix $\rho_n$ is

$$\rho_n = \tau_n - U$$

$$= (Z_i - Z_o)(Z_i + Z_o)^{-1}$$

With the transmission and reflection matrices as defined in (17) and (18), the transmission and reflection of an incident wave as a whole can be calculated. Easier would be to have transmission and reflection coefficients for each separate propagation mode. This is, however, not possible: the reflection of an incident wave depends on the total wave. If the transmission line is not properly terminated for every mode, an incident propagation mode may be reflected in an other mode. Modes that didn't exist before, can suddenly be present in the reflected wave, and these modes are usually not the desired ones.

The reflected wave $V_{bl}$ is a backward wave, propagating towards the beginning of the transmission line. When it arrives there, it is reflected again. Thus, the transmitted wave is reflected back and forth between the two terminations. A summation of all reflections results in a standing wave on the transmission line. The current associated with this standing wave is the current distribution that has to be calculated.
3.5. **Calculation of the radiated Electro-Magnetic Field.**

In the previous paragraph is described, how the current distribution along a transmission line can be calculated. When the current distribution is known, it can be used to determine the radiation of the transmission line. In this paragraph, the radiation of the two-wire model is calculated.

The two-wire model is essentially a pair of straight wires running parallel above a perfectly conducting ground plane. In [21,32], equations (19,20,21) can be found, that describe the electric and magnetic field components of a straight wire source. The current variation is assumed harmonic.

\[
\begin{align*}
H_\phi &= \frac{llk^2}{4\pi}e^{-jkr}\left(-\frac{1}{j(kr)} + \frac{1}{(kr)^2}\right)\sin\theta \ A/m \ (19) \\
E_r &= \frac{llk^3}{2\pi\omega e}e^{-jkr}\left(-\frac{1}{(kr)^2} + \frac{1}{j(kr)^3}\right)\cos\theta \ V/m \ (20) \\
E_\phi &= \frac{llk^3}{4\pi\omega e}e^{-jkr}\left(-\frac{1}{j(kr)} + \frac{1}{(kr)^2} + \frac{1}{j(kr)^3}\right)\sin\theta \ V/m \ (21)
\end{align*}
\]

Here,

\( l \) = wire current,
\( \ell \) = wire length,
\( k = \omega/c \),
\( r \) = distance from source to observation point.

The definition of the spherical coordinates \( \theta \) and \( \phi \) is shown in figure 3.5.
In equations (19,20,21), assumptions are made that don't hold on the two-wire model: the source has to have an uniform current distribution, and the distance between source and observation point has to be large compared to the length of the wire. The use of the TLM method was based on the fact that the current on the two-wire model can not be considered uniformly distributed. The second assumption is also too restrictive. With a typical length of a S-bus of 100 meters, the distance to the observation point should be a few kilometers. We are, however, primarily interested in the radiation at a few meters, which is an appropriate distance for eavesdropping the S-bus.

Nevertheless, equations (19,20,21) can be used to calculate the radiation of configurations with a non-uniform current distribution. For this the configuration has to be divided into a large number of differential sections. If the sections are small enough for the two assumptions to hold, the equations describe the contribution of each section to the EM-field. The total EM-field can then be computed by summing the contributions of all differential sections. The wire currents of the sections differ in phase. These phase differences have to be accounted for in the summation. This is done by representing the wire current $I$ as a complex value.

**Figure 3.5:** Definition of spherical coordinates for straight wire source.
The short distance of the observation point presents an other problem. In most studies of antenna problems, the distance between all point along the radiating wire and the observation point is assumed equal. The propagation delay will then also be equal along the wire. In the S-bus configuration, differences in propagation delay will have to be accounted for. The propagation delay of each field contribution must be calculated. From this propagation delay the appropriate phase shift can be calculated. In equations (19,20,21), the phase shift is represented by the term $e^{-jk_0r}$.

Each of the two wires of the two-wire model can easily be divided into small differential sections. With these sections and the current distribution on each line, the radiation of the two-wire model could be calculated. In this way, both the differential-mode and the common-mode radiation are calculated. But, on the S-bus, only the common-mode radiation is of interest. In reality, the two-wire model is twisted, which greatly reduces the differential-mode radiation. Therefore, the differential-mode radiation and currents can be disregarded. Only the common-mode current has to be considered. Because the cross-sectional dimensions of the two-wire model are small compared to the wavelength, the common-mode current can be thought to flow through one conductor at the center of the two-wire model. In this way, the two-wire model is simplified into one conductor above a ground plane, carrying only the common-mode current.

The common-mode current returns through the ground plane. So, also the contribution to the EM-field of the current in the ground field has to be calculated. In principle, this could be done in the same way as for a straight wire: dividing the ground plane into small sections with constant current and summing the contributions to the EM-field of all sections. However, the current distribution should be calculated in every point of the ground plane, which is a very complex problem. Much more convenient is to represent the ground plane by the mirror image of the conductor
[12]. Thus, a configuration of two conductors, as shown in figure 3.6., is obtained. The radiation of this configuration is easily calculated by dividing each conductor in differential straight wire sections.

\[ r_2 = \sqrt{x_p^2 + (y + H)^2 + (z - z_2)^2} \]

\[ r_1 = \sqrt{x_p^2 + (y - H)^2 + (z - z_1)^2} \]

Figure 3.6: Radiating wire configuration.

Both conductors in figure 3.6. contribute to the EM-field. Adding the respective contributions is difficult in the spherical reference frame of equations (19,20,21), because for the contribution of each differential section a different reference frame is used. Therefore, a Cartesian reference frame is defined, in which all field components can be expressed. The spherical coordinates in equations (19,20,21) have to be converted to these Cartesian coordinates. For this the following relations are used:
\[
\begin{align*}
\cos \theta_1 &= \frac{z_p - z_1}{r_1} & \quad \sin \theta_1 &= \frac{\sqrt{x_p^2 + (y_p - H)^2}}{r_1} \\
\cos \theta_2 &= \frac{z_p - z_1}{r_2} & \quad \sin \theta_2 &= \frac{\sqrt{x_p^2 + (y_p + H)^2}}{r_2} \\
\cos \phi_1 &= \frac{x_p}{\sqrt{x_p^2 + (y_p - H)^2}} & \quad \sin \phi_1 &= \frac{x_p}{\sqrt{x_p^2 + (y_p - H)^2}} \\
\cos \phi_2 &= \frac{x_p}{\sqrt{x_p^2 + (y_p + H)^2}} & \quad \sin \phi_2 &= \frac{x_p}{\sqrt{x_p^2 + (y_p + H)^2}}
\end{align*}
\]

where the suffix \( 1 \) or \( 2 \) denotes the conductor, to which the spherical coordinates relate. The definition of the Cartesian coordinates is shown in figure 3.6.

The field components in equations (19,20,21) have to be converted too. In the Cartesian reference, the field components become:

\[
\begin{align*}
H_x &= H_\phi \cdot \cos \phi \\
H_y &= H_\phi \cdot \sin \phi \\
E_x &= E_r \cdot \sin \theta \sin \phi + E_\theta \cdot \cos \theta \sin \phi \\
E_y &= E_r \cdot \sin \theta \cos \phi + E_\theta \cdot \cos \theta \sin \phi \\
E_z &= E_r \cdot \cos \theta - E_\theta \cdot \sin \theta
\end{align*}
\]

After conversion of the field components and the spherical coordinates \( \theta \) and \( \phi \), the contribution of conductor 1 to the electrical and magnetic field can be written as:
\[ H_x = \frac{1}{4\pi} \frac{1}{1} \left( \frac{1}{jkr} + \frac{1}{(kr)^2} \right) y_p - H \]

\[ H_y = \frac{1}{4\pi} \frac{1}{1} \left( \frac{1}{jkr} + \frac{1}{(kr)^2} \right) x_p \]

\[ E_x = \frac{1}{4\pi \omega c} \frac{1}{1} \left( \frac{1}{jkr} + \frac{3}{(kr)^2} + \frac{3}{j(kr)^3} \right) \frac{(z - z_i) x_p}{r^2} \]

\[ E_y = \frac{1}{4\pi \omega c} \frac{1}{1} \left( \frac{1}{jkr} + \frac{3}{(kr)^2} + \frac{3}{j(kr)^3} \right) \frac{(z - z_i)(y_p - H)}{r^2} \]

\[ E_z = \frac{1}{4\pi \omega c} \frac{1}{1} \left\{ \frac{2}{(kr)^2} + \frac{2}{j(kr)^3} \right\} \left\{ \frac{-1}{jkr} + \frac{3}{(kr)^2} + \frac{3}{j(kr)^3} \right\} \frac{x_p^2 + (y_p - H)^2}{r^2} \]

with \( r = r_i \). For the contribution of conductor 2, a similar expression is obtained, with \( r = r_z \) and with \( (y_p + H) \) instead of \( (y_p - H) \).

3.6. Software implementation.

In the previous paragraph, the total EM-field is shown to be formed by a summation of field contributions of currents on a large number of differential sections. The current on each section is itself a summation of reflections. If the differential sections are made infinitely small, the summation of field contributions can be represented by an integral. This integral is very complex, even if the current is equally distributed. Especially with a non-uniform current, it is far too complicated for an analytical solution. Therefore, the EM-field is calculated numerically on a computer.
For the numerical computation, a set of two Pascal programs is implemented. With the first program, called R9r50M, the common-mode current is calculated. The second program, called R4D, is used to compute the resulting EM-field. The source code of both programs is given in Appendix A and B; in this paragraph, a brief description of the programs is given.

The program R9r50M computes the common-mode current in a number of points along the transmission line. It generates two files: a configuration file (with extension .CON) in which the configuration parameters are recorded, and a data file (with extension .DAT) in which the common-mode current in each point is written. The data file is the input file for the R4D program.

When R9r50M is started, it first asks for configuration parameters, such as diameter and height, of the transmission line. With these parameters the L and C matrices are calculated. The C matrix is computed by calculating the mutual capacitances and capacitances to ground of all conductors. The mutual capacitance $C_{ij}$ between conductor $i$ and $j$ is described by a very complex expression [32]:

$$C_{ij} = -\frac{2\pi \varepsilon_{\text{eff}} \varepsilon_0}{\text{Det}} \cdot P_{12}$$

where,

$$P_{12} = \text{arccosh} \left( \frac{D}{2a} \right) - \text{arccosh} \left( \frac{\sqrt{h_1 h_2 + D^2}}{2a} \right)$$

$$\text{Det} = \text{arccosh} \left( \frac{h_1}{a} \right) \cdot \text{arccosh} \left( \frac{h_2}{a} \right) - P_{12}^2$$

(27)
\[ \varepsilon_{\text{eff.}} = 1 + \left( \frac{b}{a} \right)^2 - 1 \cdot \left\{ \varepsilon - 1 \right\} \]

and

\[ a = \text{radius of conductor} \]
\[ b = \text{radius of insulation} \]
\[ D = \text{wire separation} \]
\[ h_1 = \text{height of conductor 1} \]
\[ h_2 = \text{height of conductor 2} \]

The capacitance to ground \( C_{11} \) of conductor 1, is described by the expression:

\[ C_{11} = \frac{2\pi \varepsilon_0}{\text{Det}} \arccosh \left( \frac{h_2}{a} \right) \]

with \( \text{Det} \) from equation (27).

Because no sufficiently accurate equations for mutual and self inductance could be found, the inductance matrix \( L \) is calculated in an entirely different way. As suggested in [24], the matrix \( L \) can be derived from matrix \( C \) if the dielectric is homogeneous. In that case, both propagation modes coincide, and equation (5) reduces to \( LC = (1/\psi_o^2)U \). So, if \( C_a \) is computed in absence of wire insulation, \( L \) can be calculated using

\[ L = \frac{1}{\psi_o^2} C_a^{-1} \]

with propagation velocity \( \psi_o = 1/\sqrt{\mu \varepsilon} \) is the speed of light.

After the \( L \) and \( C \) matrices are thus calculated, they are used to compute the propagation velocities and modes. The program then calculates the starting forward wave, and splits it into propaga-
tion modes. In each point along the line, the phase shifts for both modes are calculated, and the propagation modes, each with its appropriate phase shift, are added. The two propagation modes together determine the current vector. For each point this current vector is stored in a current-array. Thus, the contribution to the current distribution of the starting forward wave is computed. Now the first reflection is calculated, and its contribution to the current distribution computed and added in the current-array. After a number of reflections the propagating wave becomes too small to significantly change the current distribution: further reflections can be disregarded. The calculation of the current distribution is completed, and the common-mode currents can be written to the data file.

In the program RIECOM, one deviation from the TLM theory is incorporated. The TLM theory assumes a lossless transmission line; a real transmission line, however, is not lossless. Dissipation in the conductors and radiation both cause a loss of energy. Though these losses are only small, they may be of influence, if the transmission line is long. As part of the configuration of the transmission line, a loss ratio can be given to RIECOM in dBs per kilometer. The way in which this loss is implemented describes dissipation of energy. Radiation of energy is too difficult to incorporate. The advantage of dissipation is that it is constant along the transmission line: a propagating wave loses a constant part of its energy for every meter it propagates. In RIECOM, for each point, together with the phase shift, a loss ratio relative to the distance the wave has travelled is calculated. The current vector is multiplied with this ratio, before it is added in the current-array.

The radiation loss is relative to the standing wave not to the propagating wave. It therefore varies along the transmission line. To calculate the radiation loss, first the current distribution has to be calculated. From this current distribution the radiated
energy from each point can be computed. But the radiation influences the current distribution. The current distribution has to be recalculated, and a new radiation loss must be computed. This process has to be repeated, until the current distribution doesn't change anymore. Because of the complicated procedure involved, in RADOM no adaptation to the current distribution is made for radiation losses.

The program RAD is much simpler of conception. It calculates the EM-field in a number of points along a straight observation line. After the number of points and the observation line are given, a loop of observation points is started. In each observation point the EM-field components (with appropriate phase shifts) of all differential sections along the transmission line are calculated and summed. The total EM-field is then written to the output file: the magnetic field in its \(H_x\) and \(H_y\) components, the electrical field in its \(E_x\), \(E_y\), and \(E_z\) components.
4. VERIFICATION OF THE MODEL.

4.1. Objectives of verification.

In chapter three, the modelling of the S-bus in a two-conductor model is described. With the programs RJECOM and RAD the common-mode radiation of this two-conductor model can be calculated using TLM theory. In this chapter, an attempt is made to verify the accuracy of these programs. Only a simple test to validate the calculated results was conducted; a full verification would entitle a search for the range of coverage of the theory, for which neither was the time nor were the technical possibilities available.

Even in a simple compliance test, a number of problems is encountered. Most of these problems are related to space requirements. The objective of the use of TLM theory was to predict wavelength effects. Therefore, wavelength effect should also be included in the test, requiring a transmission line length of at least a half wavelength. At frequencies of interest in the S-bus, a half wavelength amounts to approximately 100 meters. For accurate EM-field measurements, not only in the direction of the transmission line, but also in other directions a large space is required, with a conducting ground and free from other metal structures. The space requirements can be considerably reduced by scaling up the frequency: at 100 MHz a transmission line of a few meters would suffice to measure wavelength effects. But even at 100 MHz, a very large conducting ground plane is required for EM-field measurements. Scaling up the frequency much further is impractical, because other problems such as parasitic capacitances and inductances would hamper the accuracy of the measurement.

Unable to meet the requirements for EM-field measurements, it was decided to restrict the test to the measurement of the current distribution on the transmission line. In this way, only the accuracy of RJECOM could be tested. But, since in RJECOM most of the
assumptions are made that could hamper the validity of the calculated radiation, this was considered a reasonable compromise.

On a two-conductor transmission line, both the common-mode and differential-mode currents can be measured. With the program RJECOM, only common-mode currents can be calculated. Therefore, a few alterations were made in RJECOM to get a program EURP, which computes differential-mode currents. Now the differential-mode as well as the common-mode currents can be compared with measured results, to get a better test of the TLM theory. Another way in which the reliability of the test can be increased is variation of the terminations of the transmission line and of the cable symmetry. By varying the terminations, the amount of reflection is changed, an effect which the TLM theory certainly has to predict. A change in cable symmetry effects the propagation modes on the cable, and will increase the common-mode current.

In the verification test, small common-mode currents are measured. These common-mode currents, even if they are small, can cause considerable radiation, because of the relatively large common-mode loop. The large loop is, by reciprocity, also the cause of the great susceptibility to irradiation of common-mode current measurements. Especially in these measurements, it is important to pay attention to EMC aspects of the measurement set-up.

4.2. The measurement set-up.

For the verification test, a 4.75 meter long two-conductor transmission line was built. With a frequency of 50 MHz, this ensured wavelength effects to be present. The transmission line was suspended at a height of 20 centimeter above an approximately 25 centimeter wide strip of aluminium foil. Both conductors consisted of PVC coated wire ($\varepsilon = 3.5$), with a conductor diameter of 0.4 millimeter and an outer diameter of 0.85 millimeter. The wire separation was 10 millimeter. The conductors were positioned
using circular wooden boards with two holes, through which the conductors were stuck. The symmetry of the transmission line could be changed by turning the boards, and thus turning the plane in which the two conductors are lying. Maximum symmetry was obtained when the two conductors are in a horizontal plane, maximum asymmetry when they were in a vertical plane. The axis of the transmission line always remained in place at a height of 20 centimeter.

The 50 MHz source signal was generated by a Marconi Instruments MF/HF AM/FM signal generator type TF2002AS. The output signal of this generator is unbalanced. Because the input signal on the transmission line had to be balanced, a Radiometer balancing transformer type UBT 3a was used. The balancing transformer has an input impedance of 75 Ω, while the output impedance of the signal generator is 50 Ω. A impedance matching pad was used to prevent reflections caused by this mismatch.

The current distributions on the transmission line were measured with four different terminations at the end of the transmission line. The termination at the beginning of the transmission line (greatly determined by the output impedance of the balancing transformer) was not changed. For maximal reflection, a short circuit and an open circuit termination were used. For minimum reflection, an approximately matched termination was build. The fourth termination was build to resemble the termination at the beginning of the transmission line. Both the matched termination and the termination resembling the beginning of the transmission line (terminated end) were made of resistors. With a HP 4815A vector impedance meter, the impedances of these terminations were measured as good as possible. The impedances of the termination at the beginning of the transmission line were also measured. In table 4.1. the results of these measurements are stated. No impedances could be measured for the open and short circuit end terminations.
To improve the reliability of the test, not only the common-mode, but also the differential-mode current distribution was measured. Differential-mode currents were measured on a symmetric transmission line, common-mode currents on an asymmetric line, to increase the current amplitude. Currents were measured at 25 centimeter intervals.

The differential current was measured with a Tektronix P6020 AC current probe on conductor 1. The current probe was connected to a HP 8405A vector voltmeter to read out the measured current. The vector voltmeter was locked in at 50 MHz, via a connection to the synchronization output of the signal generator.

For a measurement of the common-mode current, the current probe must encircle both conductors. The P6020 current probe was not big enough for that. In theory, it should be possible to measure the currents on both conductors and then subtract them (vector calculation) to get the common-mode current. The differential-mode to common-mode ratio, however, was too high to get significant results this way. Therefore, a custom-made current probe was made, that fitted around both conductors.

The principle of a current probe is based on magnetic coupling. A current probe is in fact a current transformer of which the primary winding is the conductor of the current that is measured. In the secondary windings, a current is induced, which can be
measured. To improve the magnetic coupling ferrite material is used.

The custom-made current probe used in this test is made of a ferrite ring with a high relative permeability at high frequencies. The ferrite material, put around the transmission line, concentrates the magnetic flux in the ring. A secondary winding is wound around the ring to pick up this magnetic flux. If the magnetic permeability of the ferrite would be infinitive, all magnetic flux would be concentrated in the ring. The magnetic coupling to the secondary winding would then be insensitive to the exact position of the current probe. The relative permeability, however, is not that high, and the current probe should be carefully positioned. Important is that the coupling of both conductors is kept equal.

If no extra precautions are taken, the current probe exhibits not only magnetic but also capacitive coupling. This capacitive coupling, though increasing the total coupling, is highly undesirable. First, it is not current dependant but voltage dependant, and the voltage distribution need not be in phase with the current distribution. Furthermore, the capacitive coupling is very sensitive to the positioning of the current probe, since the electric field is not concentrated by the ferrite material. To diminish the effect of capacitive coupling, the secondary windings were shielded with an piece of copper foil. The copper foil was laid in an open winding around the other windings, with one of the ends earthed.

Since the common-mode current is much smaller than the differential-mode current, a greater sensitivity is required. Also a greater selectivity is needed, because irradiation of signals at other frequencies create common-mode currents which would otherwise be measured too. The HP vector voltmeter turned out to be not sensitive enough to measure the common-mode current. Therefore, a
HP 8552A/8553B wave analyzer, combining a high sensitivity with a high selectivity, was used.

A disadvantage of the custom-made current probe was that it had to be calibrated. This was done with the Tektronix current probe. The current in a wire connected to the signal generator was measured, both with the custom-made current probe plus wave analyzer, and with the Tektronix current probe plus vector voltmeter. A series of measurements with differential current magnitude was carried out, all at a frequency of 50 MHz. The sensitivity of the custom-made current probe turned out to be 0.272 μA/μV.

4.3. Comparing the computed and measured current distribution.

For a comparison with the measured results, the current distributions had to be computed with RICOM and EURY. It is most important that the transmission line parameters, on which RICOM and EURY base their calculations, are entered correctly: the configuration used in the calculations has to represent the actual configuration. The physical dimensions of the transmission line (height, length, diameters etc) were measured as good as possible from the actual configuration. For the termination at the beginning of the line and for the matched circuit and terminated end terminations, the measured impedance values from table 4.1. were entered. Since the impedances for the open circuit and short circuit end terminations could not be measured, somewhat arbitrary values had to be used. For a short circuit end termination, the amplitude as well as the argument should be zero. Zero impedances, however, caused division by zero errors in RICOM and EURY. Because the impedance in the actual termination would neither be exactly zero, it was decided to enter an amplitude of 0.001 Ω and a zero argument. The impedances in the open circuit end termination were even more arbitrary. Because of parasitic capacitances, they would not be infinitive. Therefore, impedances of 100 kΩ with arguments of −90° (capacitance) were assumed.
A more difficult problem was to determine the magnitude of the input signal. To exclude the influence of the balancing transformer, this signal should be measured after the transformer. It was, however, impossible to measure the output voltage of the balancing transformer, without pulling it out of balance. This problem was bypassed by measuring not the output voltage but the output current. Both the differential-mode and the common-mode output current were measured, each with disconnected transmission line to prevent reflections from interfering. The measured values were: 0.44 mA for the differential output current and 8 µA for the common-mode output current.

When all the necessary parameters were available, the current distributions could be computed and compared with the measured results. In figure 4.1, the computed differential-mode currents are shown together with the measured currents. The computed and measured distributions correspond very well. The greatest difference is seen at the open circuit end termination, with a difference of approximately 3 dB. The wavelength of the measured current distributions is somewhat shorter than that of the computed distributions, indicating a lower propagation velocity.
Figure 4.1.: Differential-mode current distributions: (a) Short circuit, (b) open circuit, (c) matched circuit, and (d) terminated end termination.
The computed common-mode current distributions are somewhat more erratic, as shown in figure 4.2. Inaccuracy in the measurement of the common-mode current could be responsible for the difference in amplitude at the short circuit and terminated end terminations. A possible explanation for the phase shifts between computed and measured distributions at the matched circuit and open circuit terminations could be inaccuracies in the measured common-mode impedances.

All in all, the computed and measured results correspond fairly well; especially if the differences are seen in dB's, which is a more common measure in EMC surroundings. The programs RIGCOM and GURY seem to be reliable. At least, no proof of their unreliability is given.
Figure 4.2: Common-mode current distributions: (a) Short circuit, (b) open circuit, (c) matched circuit, and (d) terminated end termination.
5. INFLUENCE OF CONFIGURATION PARAMETERS.

5.1. Effects on the current distribution.

The purpose of the two programs $\text{RJECA}$ and $\text{RAD}$ is to investigate the influence of configuration parameters on the EM-field. Which parameters need most attention to build a transmission line with low susceptibility to eavesdropping, and which parameters can be disregarded. Primary, we are interested in the effects of the configuration parameters on the EM-field, but since the EM-field is determined by the common-mode current, effects on the current distribution should also be investigated. In this paragraph, attention is focused on the effect on the current distribution; in the next paragraph, the effects on the EM-field are discussed.

The influence of a configuration parameter can be determined by calculating the effect of parameter changes. Starting from a basic configuration, one of the configuration parameters is changed, and the resulting current distribution is computed. The basic configuration used here was a 150 meter long transmission line at a height of 0.5 meter. The conductors, with a inner diameter of 0.5 millimeter and an outer diameter of 0.9 millimeter, resembled the conductors in an actual telephone cable. The transmitted signal was a symmetric sine wave with an amplitude of 0.75 V and a frequency of 1 MHz. On both ends the transmission line was terminated in a matched termination.
The first parameter that was changed was the output unbalance of the transmitter. In figure 5.1., the current distribution is shown for output unbalances of 5\% and 10\%. The output unbalance is the most important cause of common-mode current. In this case, an 5\% unbalance already causes a common-mode current of ± 60 dB \(\mu A\). The current distribution is linearly dependent on the output balance. An 6 dB increase in unbalance results in an equal increase of the common-mode current.

*figure 5.1.*: Effect on current distribution of transmitter output unbalance (high impedance end termination).
Another cause of common-mode current is the unbalance of the common-mode impedances in the terminations. In figure 5.2., the effect of two different unbalances in an end termination is shown. High impedances were used with unbalances of 5% and 10%. The common-mode current is linearly dependent on the percentage of unbalance. The influence of the unbalance in terminations is less than the influence of output unbalance: where a output balance of 5% caused a common-mode current of ±60 dB μA, the common-mode current for a 5% unbalance in the end termination is only ±25 dB μA. The influence of termination unbalance increases when there are more reflections, because each time a wave is reflected at an unbalanced termination, an extra common-mode component is introduced. In the configuration used to get figure 5.2., the termination at the beginning of the transmission line was matched, so only one reflection existed.

Cable asymmetry is the last cause of common-mode current. Of the three parameters causing common-mode currents, it is least effective. In figure 5.3., where the maximum cable asymmetry is used, can be seen that the resulting common-mode current is only
-20 dB µA. It is, however, uncertain if the cable asymmetry is large enough to represent actual cable unbalances. In the two-conductor model, the cable unbalance is due to a difference in height of the conductors, and is, therefore, limited by the conductor spacing.

\[
\begin{array}{c|c|c|c|c}
\text{Position in meters} & 0 & 50 & 100 & 150 \\
\text{Current in µA} & 0.00 & 0.05 & 0.10 & 0.15 & 0.20 \\
\end{array}
\]

\textbf{figure 5.3.: Effect on current distribution of cable unbalance (high impedance end termination).}

Besides parameters that cause common-mode currents, there are also parameters that only influence the magnitude of the common-mode current. These parameters do not cause common-mode currents and are, therefore, only important if common-mode currents already exist. Nevertheless, they should be investigated. One of these parameters is reflection. If any of the terminations is not matched, waves propagating towards it are partly reflected. No independent reflection coefficients for both propagation modes exist, but the amount of reflection is generally different for each mode. The reflection of the common-mode wave, which is most important in this context, is determined by the common-mode impedances.
In figure 5.4.(a), the current distribution is shown for a four different common-mode impedances in the end termination. The common-mode current is inserted on the transmission line by an output unbalance. If the common-mode impedances are matched, there is no reflection, and a constant current distribution results. A termination in which the common-mode impedances are short circuits or open circuits causes maximum reflection. The forward propagating common-mode wave is totally reflected toward the beginning of the transmission line. Because of the matched termination at the beginning of the transmission line, the wave is not reflected any further. A short circuit permits high currents, therefore the maximum of the current distribution is at the end of the line. With open circuits in the end termination, the common-mode current is forced zero. Since the length of the transmission line is a halve wavelength, this results in an current distribution which is exactly one sine lobe. That not only the magnitude of the impedance is important, is shown in the last curve of figure 5.4.(a): if an impedance is used with a matching magnitude but with a non-zero argument, the propagating wave is still reflected.
figure 5.4: Effect on current distribution of reflections:
(a) \( f = 1 \text{ MHz} \), (b) \( f = 0.1 \text{ MHz} \).

Looking at the curves in figure 5.4(a), we see that reflections not only cause a local rise in the current distribution; the average common-mode current is also increased. At lower frequencies, the effect on the average common-mode current is even
more pronounced. In figure 5.4.(b) is illustrated, that a high impedance in the end termination not only forces down the current at the end, but along the whole transmission line. A low common-mode impedance, on the contrary, results in an increase in current.

Waves can also be reflected at the termination at the begin of the transmission line. So impedances in that termination are of interest. There is, however, also an other mechanism in which these impedances influence the common-mode current. If the transmitter is not perfectly balanced, it generates a common-mode voltage. The common-mode output impedance determines the amount of common-mode current caused by this voltage. This effect is illustrated in figure 5.5.

\[ \text{figure 5.5.: Effect on current distribution of output impedance.} \]

The last parameter that influences the amount of common-mode current is the conductor loss. The effect of loss is shown in figure 5.6. Conductor loss is proportional to the propagating wave not to the standing wave. It can, therefore, also cause a local increase in current, as seen in figure 5.6.
The computations of the parameter effects on the current distribution show that the most important cause of common-mode current is the output unbalance. It is obvious that the output unbalance should be kept as low as possible. More interesting is the question which termination is best. Should the common-mode impedances be matched to prevent reflections, or should high impedances be used to keep the current low. If the transmission line has a length of a half wavelength or a multiple of that length, high impedances can result in very high common-mode currents. In other situations, however, they force down the common-mode current. The higher the impedances; the more pronounced the effects. Furthermore, higher impedances increase the frequency selectivity of the resonance effects.
5.2. Effects on the Electro-Magnetic field.

In this paragraph, the EM-field of a basic configuration is calculated. Various parameters of the configuration are then changed to determine their effect on the EM-field. As basic configuration, the same 150 meter transmission line as in the previous paragraph is used. A common-mode current is introduced by an 10 % unbalance of the transmitter output. At both ends of the transmission line, matched terminations are assumed, to get a constant current distribution. The excitation frequency is 1 MHz. EM-field versus distance curves were calculated along a 0.5 meter high observation line at the middle of the transmission line. Distances varied over a range of interest from 10 centimeter to 100 meter. Larger distances are not important for eavesdropping purposes, because the EM-fields will be too small at that distance and because other S-busses will probably interfere.

Before the EM-field of the basic configuration can be computed, a closer look has to be taken at the number of points in which the current distribution is calculated. The accuracy with which RAD computes the EM-field is greatly dependent on the number of points in the current distribution file. At large distances, only a few points are needed, but it is impossible to accurately calculate the EM-field at 1 centimeter, if the current distribution is known only every other meter. Because 10 centimeter is about the closest distance of interest, the current distribution was calculated every 5 centimeter. A large number of points, however, yields very large input files, and, more serious, needs very much computing time. If matched terminations are used (few reflections), the computing time of RICOM is less affected by the large number of points than RAD. Therefore, the EM-field is only at close range computed with one points per 5 centimeter. At distances larger than 1 meter, a current distribution every 15 centimeter was used, even though this meant that the current distribution had to be calculated twice.
With a frequency of 1 MHz, the range of 0.1 meter to 100 meter falls in the nearby field region. In this region, the EM-field can be expected to decrease proportionally to \(1/r^2\) for magnetic field components and \(1/r^3\) for electric field components. Looking at figure 5.7., however, we see that the computed EM-field versus distance curves are quite different from expected. Where a decrease of 40 dB/dec and 60 dB/dec was expected, the actual decrease is much less. In the 10 meter to 100 meter range, gradually a fall rate of 40 dB/dec for \(H_x\) and \(E_y\) components and 60 dB/dec for \(H_y\) and \(E_x\) components is reached. But in the 10 centimeter to 1 meter range, the \(H_x\) and \(E_y\) components are almost constant at ±40 dB \(\mu A/m\) and ±95 dB \(\mu A/m\) respectively. The \(H_y\) and \(E_x\) components only drop approximately 20 dB. The \(E_z\) component decreases very much over the first meter. Only in the first 40 centimeter it is of any importance. At a distance of 1 meter it is approximately 120 dB below the other electric field components. The rapid decrease of the \(E_z\) component implies that \(E_z\) is of no importance to the susceptibility of the transmission line to eavesdropping. The \(E_z\) component is, therefore, not included in any of the other EM-field curves in this paragraph.
There are a number of configuration parameters that could influence the EM-field. Most important of these is the magnitude of the current distribution. The influence of the current distribution is obvious. From equation (19,20,21), we easily see that the EM-fields are linearly dependent on the current magnitude. An overall increase in current along the transmission line results in an equal overall increase of the EM-field components. Because of the direct dependence of the EM-fields on the current magnitude, it is important to keep the currents low.

Another parameter that could influence the EM-field is the amount of reflection. Reflections also cause an increase in current, only this increase is more local and is accompanied with a decrease in current elsewhere. Two calculations were made to determine the effect of reflections: one with very small common-mode impedances in the end termination and one with very large common-mode impedances. The termination at the beginning of the transmission line was kept matched. In both situations a maximum reflection is obtained at the end of the line, but the resulting standing wave patterns differ in phase. In figure 5.8. (a), EM-field curves are shown for the high impedance termination. With a high impedance termination, the current is at its maximum at the middle of the transmission line. The local 6 dB increase in current results in an 6 dB increase in the magnetic field components. The electric field components, however, drop dramatically with ±75 dB. In figure 5.8. (b) is shown that these are only local effects.
With a low impedance termination reverse effects are obtained. Here the current is at its minimum at the middle of the transmission line, resulting in a smaller magnetic field. In figure 5.9. is shown that the magnetic field increases again in
the 10 meter to 100 meter range. This increase is probably caused by the higher currents at both ends of the transmission line.

\[ H_x - H_y \]

\[ E_x \]

\[ E_y \]

\[ \text{Distance in meters} \]

**Figure 5.9:** Effects on EM-field of reflections: low common-mode impedance termination.

The presence of reflections can locally cause a significant rise in local EM-field components. If the right place is found, an eavesdropper could profit from this. However, each frequency will have its own "right places" and "wrong places". There is probably no place along the transmission line where reflections cause an increase in the EM-field for all frequencies of interest in the S-bus signal.

The third factor of influence on the EM-field is the frequency. Generally, in literature on EMC problems is said that radiation increases with increasing frequency. Equations (19,20,21) show, however, that this is not true in the nearby field region. In figures 5.10.(a) and (b), where the EM field components are plotted for frequencies of 0.1 MHz and 10 MHz, can be seen that a change in frequency only affects the EM-field at distances of more than 10 meter. The frequency range of 0.1 MHz to 10 MHz covers most of the frequencies of interest in the S-bus signal. Only if the frequency is increased much further, will it influence the
EM-field at shorter distances.

\[ \text{Distance in meters} \]

(a)

Hx - - Hy — Ex — Ey

Distance in meters

(b)

Hx - - Hy — Ex — Ey

Distance in meters

Figure 5.10: Effect on EM-field of frequency: (a) \( f = 0.1 \text{ MHz} \), (b) \( f = 10 \text{ MHz} \).

The height and the length of the transmission line are the last parameters that could have an effect on the EM field. Both parameters determine the size of the common-mode loop. If the transmission line would be electrically short, the EM field would
be directly proportional to the area enclosed in the common-mode loop. Figure 5.11., where the height of the transmission line is increased to 1 meter, and figure 5.12.(a) and (b), where the EM fields are plotted for 15 meter and 1500 meter long transmission lines, show that the matter is more complicated for electrically long lines. The field components of the 15 meter line are indeed smaller than those of the 150 meter line; but only at distances more than a couple of meters. The effect is greater on the electric field than on the magnetic field. The longer transmission line of 1500 meter also has got a smaller EM-field than the 150 meter line, contrary to what was expected. Because the currents on the transmission line differ in phase, their contributions can not simply be added. A longer transmission line can, therefore, have a smaller EM-field. The field curves for the transmission line with increased height in figure 5.11. show, at greater distance, the increase that was expected. More important is the increase of the region in which $H_x$ and $E_y$ are constant.

*figure 5.11.: Effect on EM-field of transmission line height: $h = 1$ m.*
Figure 5.12: Effect on EM-field of transmission line length: (a) $L = 15$ m, (b) $L = 1500$ m.
It can be concluded, from the EM-field computations, that the EM-field is quite insensitive to most of the configuration parameters. Some effect of parameter changes can be seen, but no dramatic change in the EM-field curves is resulting. Only changes in the current magnitude and reflections produce significant effects. The current magnitude is of overall importance; reflections produce only local effects. Moreover, the effects of reflections are frequency dependent. Over a wider frequency spectrum, they cause no greater susceptibility to eavesdropping.
6. CONCLUSIONS AND RECOMMENDATIONS.

Previous research [5, 6, 7, 8, 26] has shown that the radiation of the S-bus presents a serious problem. Therefore more attention should be paid to the EMI aspects of the S-bus. In the CCITT Recommendation 4.30, were the Layer 1 characteristics of the S-bus are specified, these EMI aspects are neglected. Parameters that are important in this context are not or insufficiently specified. Worst case configurations within the S-bus specifications will almost certainly have a sufficiently high radiation level to make eavesdropping possible. It would be desirable to change the CCITT recommendations with respect to the parameters of concern in the EMI context. But, since the recommendations have become well established internationally, making changes would be very difficult. A more probable solution will be to adapt the PTT specifications of S-bus installations. More research is needed to determine, if it will be possible to specify a S-bus with sufficiently low radiation levels, operating on conventional telephone wires. Otherwise shielded cables will have to be used.

The radiation of the S-bus is predominantly determined by the common-mode current. Therefore, when defining S-bus specifications, most attention should be paid to the common-mode circuit. The most important cause of common-mode current is the output unbalance of the transmitter. Limiting the output unbalance would, however, require a change in the CCITT recommendations. An important parameter, that does not cause common-mode current but influences its magnitude, is the common-mode impedance of the terminations. These impedances are more easily specified in the PTT regulations, without changing CCITT recommendations. A choice should be made between high common-mode impedances and matched common-mode impedances. High impedances force down the common-mode current at most frequencies. But if resonances occur, high common-mode impedances result in very high common-mode currents.
Using matched common-mode impedances prevents resonances, but yields a higher common-mode current at all other frequencies.

Besides parameters that affect the radiation via the common-mode current, there are a number of parameters that directly influence the radiation, eg. the length and height of the transmission line. Changing these parameters, within practical S-bus situations, produces no significant effect on the EM-field patterns.
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APPENDIX A: SOURCE CODE OF RICOM.

PROGRAM ricom(input,output,cout,dout);
CONSTpi = 3.1415926536;
tpi = 6.2831853072;
pit = 1.5707963268;
eps = 8.8419412836E-12;
rt2 = 1.4142135623;
lni0 = 2.302585093;
maxitt = 1000;
maxpoint = 1000;
TYPE complex = RECORD
  r : real;
  i : real
END;
matrix = ARRAY [1..2,1..2] OF real;
cmatrix = ARRAY [1..2,1..2] OF complex;
vector = ARRAY [1..2] OF real;
cvector = ARRAY [1..2] OF complex;
cvecarr = ARRAY [0..maxpoint] OF cvector;

PROCEDURE rtoc(VAR z:complex;amp,arg:real);
(conversion real to complex)
BEGIN
  z.r:=amp*cos(arg);z.i:=amp*sin(arg)
END;

PROCEDURE mtoc(VAR z:cmatrix;amp:matrix);
(conversion real matrix to complex)
VAR i,j : integer;
BEGIN
  FOR i:=1 TO 2 DO FOR j:=1 TO 2 DO rtoc(z[i,j],amp[i,j],0)
END;

FUNCTION ca(x:complex):real; (amplitude |x| of complex)
BEGIN
  ca:=sqrt(sqr(x.r)+sqr(x.i));
END;

FUNCTION cp(x:complex):real; (argument arg(x) of complex)
VAR h : real;
BEGIN
  WITH x DO
  BEGIN
    IF r=0 THEN h:=pit
    ELSE h:=arctan(i/r);
    IF r<0 THEN BEGIN IF i<0 THEN h:=h-pi ELSE h:=h+pi END
    cp:=h
  END;
END;
FUNCTION clen(x:cvector):real;  {length of complex vector}
BEGIN
  clen := sqrt(sqr(ca(x[1])) + sqr(ca(x[2])))
END;

PROCEDURE cadd(VAR z:complex;x,y:complex);  {complex addition}
BEGIN
  z.r := x.r + y.r; z.i := x.i + y.i
END;

PROCEDURE cvadd(VAR z:cvector;x,y:cvector);
{complex vector addition}
BEGIN
  cadd(z[1],x[1],y[1]); cadd(z[2],x[2],y[2])
END;

PROCEDURE cmadd(VAR z:cmatrix;x,y:cmatrix);
{complex matrix addition}
BEGIN
  cadd(z[1,1],x[1,1],y[1,1]); cadd(z[1,2],x[1,2],y[1,2]);
  cadd(z[2,1],x[2,1],y[2,1]); cadd(z[2,2],x[2,2],y[2,2]);
END;

PROCEDURE csub(VAR z:complex;x,y:complex);
{complex subtraction}
BEGIN
  z.r := x.r - y.r; z.i := x.i - y.i
END;

PROCEDURE cmul(VAR z:complex;x,y:complex);
{complex multiplication}
BEGIN
  z.r := x.r * y.r - x.i * y.i; z.i := x.r * y.i + x.i * y.r
END;

PROCEDURE cvmul(VAR x:cvector;y:cvector;m:complex);
{multiply complex vector with number}
VAR i : integer;
BEGIN
  cmul(x[1],y[1],m); cmul(x[2],y[2],m)
END;

PROCEDURE cmmul(VAR z:cmatrix;x,y:cmatrix);
{complex matrix multiplication}
VAR k,i : integer;
BEGIN
  FOR k := 1 TO 2 DO
    FOR i := 1 TO 2 DO
      BEGIN
        cmul(u,x[k,1],y[1,i]); cmul(v,x[k,2],y[2,i]); cadd(z[k,i],u,v)
      END
    END
END;
PROCEDURE cmcmul(VAR z:cmatrix;x:cmatrix;c:real);  
{multiply complex matrix with real number}
VAR i,j : integer;
BEGIN
  FOR i:=1 TO 2 DO
    FOR j:=1 TO 2 DO
      BEGIN
        z[i,j].r:=c*x[i,j].r,z[i,j].i:=c*x[i,j].i
      END;
END;

PROCEDURE cdiv(VAR z:complex;x,y:complex);  {complex division}
VAR h : real;
BEGIN
  h:=sqr(y.r)+sqr(y.i);z.r:=(x.r*y.r+x.i*y.i)/h;
  z.i:=(x.i*y.r-x.r*y.i)/h
END;

PROCEDURE cinv(VAR y:cmatrix;x:cmatrix);  {complex matrix inversion}
VAR h,ch : complex;
BEGIN
  rtoe(y[1,1],1,0);rtoe(y[1,2],0,0);y[2,1]:=y[1,2];y[2,2]:=y[1,1];
  IF ca(x[1,1]) < ca(x[2,1]) THEN
    BEGIN
      FOR k:=1 TO 2 DO
        BEGIN
          h:=x[1,k];x[1,k]:=x[2,k];x[2,k]:=h:=y[1,k];
          y[1,k]:=y[2,k];y[2,k]:=h
        END;
    END
  END;
cdiv(h,x[2,1],x[1,1]);cmul(ch,h,x[1,2]);
csub(x[2,2],x[2,1],ch);
cmul(ch,h,y[1,1]);csub(ch,y[2,1],ch);cdiv(y[2,1],ch,x[2,2]);
cmul(ch,h,y[1,2]);csub(ch,y[2,2],ch);cdiv(y[2,2],ch,x[2,2]);
h:=x[1,2];
cmul(ch,h,y[2,1]);csub(ch,y[1,1],ch);cdiv(y[1,1],ch,x[1,1]);
cmul(ch,h,y[2,2]);csub(ch,y[1,2],ch);cdiv(y[1,2],ch,x[1,2])
END;
PROCEDURE cslv(VAR x:cvector; a:cmatrix; y:cvector);
{solve complex matrix}
VAR h, g : complex;
k : integer;
BEGIN
  IF ca(a[1,1]) < ca(a[2,1]) THEN
    BEGIN
      FOR k:=1 TO 2 DO
        BEGIN
          h:=a[1,k];a[1,k]:=a[2,k];a[2,k]:=h
          END;
          h:=y[1];y[1]:=y[2];y[2]:=h
        END;
        cdiv(h,a[2,1],a[1,1]);cmul(g,h,y[1]);csup(y[2],y[2],g);
        cmul(g,h,a[1,2]);
        csub(a[2,2],a[2,2],g);cdiv(h,a[1,2],a[2,2]);cmul(g,h,y[2]);
        csub(y[1],y[1],g);cdiv(x[1],y[1],a[1,1]);cdiv(x[2],y[2],a[2,2])
      END;
    END;
PROCEDURE func(VAR y:cvector; m:cmatrix; x:cvector);
{matrix transformation}
VAR k : integer;
h1, h2 : complex;
BEGIN
  FOR k:=1 TO 2 DO
    BEGIN
      cmul(h1,m[k,1],x[1]);cmul(h2,m[k,2],x[2]);cadd(y[k],h1,h2)
    END
  END;
PROCEDURE zpar(VAR zp:complex; z1, z2:complex);
{parallel impedance}
BEGIN
  cadd(zp, z1, z2);cdiv(zp,z2,zp);cmul(zp,z1,zp)
END;
PROCEDURE cls; {clear screen}
VAR i : integer;
BEGIN
  FOR i:=1 TO 40 DO writeln
END;
PROCEDURE yesno(VAR answer, valid:boolean);
{ask for answer yes or no}
VAR yn : char;
BEGIN
  valid:=true;readln(yn);
  IF (yn='y') OR (yn='Y') THEN answer:=true
  ELSE IF (yn='n') OR (yn='N') THEN answer:=false
    ELSE valid:=false
END;
PROCEDURE mmul(VAR z:matrix; x,y:matrix); {matrix multiplication}
VAR k,i : integer;
BEGIN
  FOR k:=1 TO 2 DO
    FOR i:=1 TO 2 DO z[k,i]:=x[k,1]*y[1,i]+x[k,2]*y[2,i]
  END;

PROCEDURE eigen(VAR vm:matrix; l,c:matrix; v:vector); {eigenvectors}
VAR ev : real;
x : matrix;
k : integer;
BEGIN
    BEGIN vm[1,1]:=l;vm[2,1]:=l;vm[1,2]:=l;vm[2,2]:=-1 END
  ELSE
    BEGIN
      mmul(x,l,c);
      FOR k:=1 TO 2 DO
        BEGIN
          ev:=1/sqr(v[k]);
          IF abs(x[2,2]-ev) > abs(x[1,2]) THEN
            BEGIN vm[1,k]:=l; v m[2,k]:=-x[2,1]/(x[2,2]-ev) END
          ELSE
            BEGIN vm[1,k]:=l; v m[2,k]:=-x[1,2]/x[1,2] END
        END
    END
  END;
END;

PROCEDURE speed(VAR v:vector; c,l:matrix); {velocity of modi}
VAR lcl,lc2,lcm,lm,cm,ll2,cl2,root,first : real;
BEGIN
  lcl:=c[1,1]*l[l,1];lc2:=c[2,2]*l[l,2];lcm:=l[l,2]*c[1,2];
  lm:=l[l,2];cm:=c[1,2];cl2:=c[1,1]*c[2,2];ll2:=l[l,1]*l[l,2];
  root:=sqrt(sqr(lcl-lc2)+4*(sqr(lm)*cl2+lcm*(lcl+lc2)+
    sqr(cm)*ll2));
  first:=lcl+lc2+2*lcm;
  v[1]:=1/sqrt((first-root)/2);v[2]:=1/sqrt((first+root)/2)
END;
PROCEDURE inv(VAR y:matrix;x:matrix);  \{matrix inversion\}
VAR  h  :  real;
    k  :  integer;
BEGIN
  y[1,1]:=1;y[1,2]:=0;y[2,1]:=0;y[2,2]:=1;
  IF abs(x[1,1]) < abs(x[2,1]) THEN
    BEGIN
      FOR k:=1 TO 2 DO
        BEGIN
          h:=x[1,k];x[1,k]:=x[2,k];x[2,k]:=h;x[1,k]:=y[1,k];y[1,k]:=y[2,k];
          y[2,k]:=h
        END
    END
  h:=x[2,1]/x[1,1];x[2,2]:=x[2,2]-h*x[1,2];
  y[2,1]:=(y[2,1]-h*y[1,1])/x[2,2];
y[2,2]:=(y[2,2]-h*y[1,2])/x[2,2];h:=x[1,2];
y[1,1]:=(y[1,1]-h*y[2,1])/x[1,1];
y[1,2]:=(y[1,2]-h*y[2,2])/x[1,1]
END;

FUNCTION  arccosh(x:real):real;
BEGIN
  arccosh:=ln(2*x)
END;

FUNCTION  cm(s,rm,h1,h2,rd,er:real):real;
\{capacitance between wires\}
VAR  ee,sa,pI2,det :  real;
BEGIN
  ee:=1+(sqr(rd/rm)-1)/(sqr((s+rd)/(2*rm))/2)*(er-1);
  sa:=sqrt(4*h1*h2+sqr(s));
  pI2:=arccosh(s/(2*rm))-arccosh(sa/(2*rm));
  det:=arccosh(h1/rm)*arccosh(h2/rm)-sqr(pI2);
  cm:=-1E-9*ee*pI2/(18*det)
END;

FUNCTION  cx(s,rm,h1,h2:real):real; \{wire to earth capacitance\}
VAR  sa,pI2,det :real;
BEGIN
  sa:=sqrt(4*h1*h2+sqr(s));
  pI2:=arccosh(s/(2*rm))-arccosh(sa/(2*rm));
  det:=arccosh(h1/rm)*arccosh(h2/rm)-sqr(pI2);
  cx:=-1E-9/(18*det)*arccosh(h2/rm)
END;

PROCEDURE  mnrn(VAR  m:matrix); \{normalize matrix\}
VAR  h  :  real;
BEGIN
  h:=sqrt(sqr(m[1,1])+sqr(m[2,1]));m[1,1]:=m[1,1]/h;
  m[2,1]:=m[2,1]/h;
  h:=sqrt(sqr(m[1,2])+sqr(m[2,2]));
  m[1,2]:=m[1,2]/h;m[2,2]:=m[2,2]/h
END;
VAR again, valid, itts, writer: boolean;
  k, j, itt, nol, noi: integer;
  c1, c2, c12, r1, r2, r1y, r2y, r1m, r2m, amp, arg, efr, ln1: real;
  posi, b, d, h1, h2, er, pv, pte, pto, wle, wlo, po, sce, scd: real;
  tev, tod, elo, elo, ple, pl0, tle, tlo, mer, mei, a: real;
  v: vector;
  c1, l, v, z, y, dv, s: matrix;
  chl, ch2, ch3, ch4, ch5, cpe, cpo, cop, con, zy1, zy2, zym: complex;
  z01, z02, z0m, z1l, zl2, zlm, zp, icom: vector;
  nulv, vie, vio, vwe, vwo: cvector;
  itot: complex;
  sym, cvm, cz, cy, y0, z0, yl, zl, m0, rm0, rml: cmatrix;
  dout, cout: text;
  filename: string;

BEGIN
  {*init*}
  again: =true; rtoc(cop, 1, 0); rtoc(con, -1, 0);
  rtoc(nulv[1], 0, 0); rtoc(nulv[2], 0, 0);
  s[1, 1]: =1; s[2, 1]: =1; s[1, 2]: =1; s[2, 2]: =-1; mtoc(sym, s);
  {*end init*}

  {*loop for more files*}
  WHILE again DO
  BEGIN

  {*ask for configuration*}
  cl;
  write('conductor radius = '); readln(a);
  write('dielectricum radius = '); readln(b);
  write('conductor spacing = '); readln(d);
  write('height conductor 1 = '); readln(h1);
  write('height conductor 2 = '); readln(h2);
  write('rel. permittivity = '); readln(er);
  writeln;
  {*end ask for configuration*}

  {*C and L matrices*}
  cl1 := cx(d, a, h1, h2); cl2 := cm(d, a, h1, h2, a, 0); c1 := cl1 + cl2;
  c[1, 1] := cl1 + cl2; c[1, 2] := c1; c[2, 1] := -cl1; c[2, 2] := -cl2;
  inv(l, c);
  FOR k := 1 TO 2 DO FOR j := 1 TO 2 DO l[k, j] := l[k, j] * 1E-16 / 9;
  cl2 := cm(d, a, h1, h2, b, er);
  c[1, 1] := cl1 + cl2; c[1, 2] := -cl1; c[2, 1] := -cl2; c[2, 2] := c1 + cl2;
  {*end C and L matrices*}

  {*propagation velocities*}
  speed(v, c, l);
  FOR k := 1 TO 2 DO dv[k, k] := v[k];
  dv[1, 2] := 0; dv[2, 1] := 0;
  {*end propagation velocities*}
(*Y, Z, and VM matrices*)

eigen(vm,1,c,v);mnrm(vm);mtodcvm, vm);

FOR k:=1 TO 2 DO BEGIN ve[k]=cmv[k,1];vo[k]=cmv[k,2] END;
inv(y,vm);mmul(y, dv, y);mmul(y,vm,y);mmul(y,c,y);
inv(z,y);
mtoc(cy,z);mtoc(cy,y);

(*matched terminations*)

ryI:=1/(y[1,1]+y[1,2]);ryZ:=1/(y[2,1]+y[2,2]);rym:=-1/y[1,2];
rzI:=z[1,1]-z[1,2];rzZ:=z[2,2]-z[2,1];rzm:=z[1,2];
rtoc(zyI,ryI,0);rtoc(zy2,ry2,0);rtoc(zym,rym,0);

writeln('matched termination: RyI = ',ryI);
writeln('RyZ = ',ryZ);
writeln('Rym = ',rym);

writeln('T matched termination: RzI = ',rzI);
writeln('RzZ = ',rzZ);
writeln('Rzm = ',rzm);

(*ask*)

writeln;
writeln('source impedance configuration (arguments/c)');
write(' :ZyI: =');readln(amp);write('arg(ZyI) = ');
readln(arg);arg:=arg-pi;rtodzOI,amp,arg);
write(':ZyZ: = ');readln(amp);write('arg(ZyZ) = ');
readln(arg);arg:=arg-pi;rtodzOZ,amp,arg);
write(':Zym: = ');readln(amp);write('arg(Zym) = ');
readln(arg);arg:=arg-pi;rtodzOm, amp, arg); writeln;

writeln('load impedance configuration');
write(' :ZyI: =');readln(amp);write(' arg(Zy1) = ');
readln(arg);arg:=arg-pi;rtoc(zll,amp,arg);
write(':ZyZ: = ');readln(amp);write('arg(ZyZ) = ');
readln(arg);arg:=arg-pi;rtodzIZ,amp,arg);
write(':Zym: = ');readln(amp);write('arg(Zym) = ');
readln(arg);arg:=arg-pi;rtoc(zIm,amp,arg); writeln;

writeln('1055 (dB/km) = ');
writeln('excitation frequency (MHz) = ');
readln(efr);

efr:=efr-Ie6;wIe:=w(1)/efr;wlo:=w(2)/efr;
writeln('wavelenght even mode = ');readln(lnl);writeln;
pte:=lnl/v[1];pto:=lnl/v[2];
writeln('propagation time even mode =',pte); writeln('propagation time odd mode =',pto);
pto:=pto*tpi*efr;pte:=pte*tpi*efr; {phase shift one transit}
write('Idiff. = ');readln(scd);scd:=scd/lOOO;
write('common mode source current in uA');
write('Icomm. = ');readln(scc);scc:=scc/lOOOOOO;
foreach:

{end *ask*}

{end *Y, Z, and VM matrices*}
(*write configuration on disk*)
writerln;
write('filename =');readln(filename);
assign(cout,filename+'.con');rewrite(cout);
writerln(cout,a); {conductor radius}
writerln(cout,b); {dielectricum radius}
writerln(cout,d); {spacing between conductors}
writerln(cout,h1); {height conductor 1}
writerln(cout,h2); {height conductor 2}
writerln(cout,er); {relative permittivity}
writerln(cout,l); {length of bus}
writerln(cout,efr); {frequency}
writerln(cout,elo); {loss}
writerln(cout,scd); {exitation diff. current}
writerln(cout,scc); {exitation comm. current}
writerln(cout,ca(zO1), ' ',cp(zO1)); {termination at z=O}
writerln(cout,ca(zO2), ' ',cp(zO2));
writerln(cout,ca(zOm), ' ',cp(zOm));
writerln(cout,ca(zll), ' ',cp(zll)); {termination at z=1}
writerln(cout,ca(zl2), ' ',cp(zl2));
writerln(cout,ca(zlm), ' ',cp(zlm));
close(cout);
@end *write configuration on disk*

(*Y, Z, reflection and transmission matrices at z=O and z=1*)
cdiv(chI,cop,zO1); cdiv(ch2,cop,zO2); cadd(yO[1,1],ch1,ch2);
cdiv(yO[1,2],con,zO1); yO[2,1]:=yO[1,2];
cdiv(chI,cop,zO2); cdiv(ch2,cop,zO1); cadd(yO[2,2],ch1,ch2);
cinv(zO,yO);
cmadd(mO,cz,zO); cinv(mO,mO); cmcmul(rmO,cz,-I);
cmadd(rmO,rmO,zO); {reflection matrix}
cdiv(chI,cop,zl1); cdiv(ch2,cop,zl2); cadd(yl[1,1],ch1,ch2);
cdiv(yl[1,2],con,zl1); yl[2,1]:=yl[1,2];
cdiv(chI,cop,zl2); cdiv(ch2,cop,zl1); cadd(yl[2,2],ch1,ch2);
cinv(zl,yl);
cmadd(mO,cz,zl); cmcmul(rml,cz,-I);
cmadd(rml,rml,zl);
cmcmul(rml,rml,mO); {reflection matrix}
@end *Y, Z, reflection and transmission matrices at z=O and z=1*

(*initialization itteration loop*)
itt:=O; itts:=true; wrl:=true;
posi:=l/lvpn; {length of differential section}
tev:=posi/v[1]*tpi*efr; tod:=posi/v[2]*tpi*efr; {phase shift per section}
ple:=exp(-ln10*elo*posi/20000); {loss per section}
tle:=exp(-ln10*elo*lnl/20000); {total loss one transit}
writerln('working');
FOR k:=O TO nopp DO itotk:=nulv;
(*begin voltage vector*)
rtodchL,scd,O);
cadd(ch2,z01,z02);cadd(ch2,ch2,z0m);cmul(ch1,ch1,ch2);
ch.r:=ca(chl);ch.l:=0;
zpar(zp,zym,z0m);
cdiv(ch3,cop,z01);cdiv(ch4,cop,zy1);cdiv(ch5,cop,zp);
cadd(ch3,ch3,ch4);
cadd(m0[1,2],ch3,ch5);cmul(m0[1,2],ch5,con);
cdiv(ch3,cop,z02);cdiv(ch4,cop,zy2);cadd(ch3,ch3,ch4);
cadd(m0[2,2],ch3,ch5);m0[2,1]:=m0[1,2];
cdiv(mm[1],chl,z0m);cmul(mm[2],con,mm[1]);cdiv(vf0,m0,mm);
rtoc(chl,ssc,0);zpar(ch2,z01,z02);cmul(ch3,ch1,ch2);
cadd(ch1,z01,zy1);cdiv(ch1,zy1,chl);cmul(ch1,ch3,ch1);
cadd(ch2,z02,zy2);cdiv(ch2,zy2,ch2);cmul(ch2,ch2,ch2);
writeln(vf0[1].r, vf0[1].i,ch1.r,ch1.i,ca(vf0[1]),ca(chl));
writeln(vf0[2].r, vf0[2].i,ch2.r,ch2.i,ca(vf0[2]),ca(ch2));
cadd(vf0[1],vf0[1],chl);cadd(vf0[2],vf0[2],ch2);
cslv(mm, cvm, vf0);cmul(vwe, ve, mm[1]);cmul(vwo, vo, mm[2]);
{split in modi}
{end *begin voltage vector*}
{end *initialize itteration loop*}

(*loop of itterations*)
WHILE itts DO
BEGIN
(*initialize section loop*)
itts:=false;
IF odd(itt) THEN
BEGIN el:=t1e;pe:=pte;po:=pto END
{begin high loss & phase shift}
ELSE
BEGIN el:=1;pe:=0;po:=0 END; {begin no loss & phase shift}
{end *initialize section loop*}

(*loop of sections*)
FOR k:=0 TO nop DO
BEGIN
(*voltage vector 1 section*)
mer:=cos(pe);mei:=-sin(pe); {shift over 1 section}
mor:=cos(po);moi:=-sin(po);
FOR j:=1 TO 2 DO
BEGIN
viel[j].r:=vwe[j].r*mer-vwe[j].i*mei;
viel[j].r:=vwo[j].r*mor-vwo[j].i*moi;
viel[j].i:=vwe[j].r*mei+vwe[j].i*mer;
viel[j].i:=vwo[j].r*moi+vwo[j].i*mor;
vitt[j].r:=(viel[j].r+viol[j].r)*el;
vitt[j].i:=(viel[j].i+viol[j].i)*el;
END;
{end *voltage vector 1 section*}
(new loss & phase shift*)
  IF odd(itt) THEN
    BEGIN pe:=pe-tev; po:=po-tod; el:=el/ple END
  ELSE
    BEGIN pe:=pe+tev; po:=po+tod; el:=el*ple END;
{end *new loss & phase shift*}

(current vector*)
  func(iitt, cy, vitt);
  IF odd(itt) THEN cvmul(iitt, litt, con);
{end *current vector*}

cvadd(itot[k], iitt, litt);
{add contribution to total section vector}
  IF clen(iitt)/clen(itot[k])*1E-20 THEN itts:=true;
{if more contribution: more itts}
END;
{end *loop of sections*}

itt:=itt+1; {next iteration}
write(itt mod 10 :1); {working indication}
IF itt= noi THEN BEGIN
  writeln;
  write('not stable, write file ? '); yesno(wri, valid);
  itts:=false {stop if enough iterations}
END;
{end *loop of iterations*}

(vector at termination and reflection*)
cpe.r:=tle*cos(pte); cpe.i:=-tle*sin(pte); cvmul(vwe, vwe, cpe);
cpo.r:=tle*cos(poto); cpo.i:=-tle*sin(poto); cvmul(vwo, vwo, cpo);
cvadd(vf, vwe, vwo); {add common + even mode}
  IF odd(itt) THEN func(vb, rm0, vf) ELSE func(vb, rm0, vf);
csvl(mm, cvm, vb); cvmul(vwe, ve, mm1); cvmul(vwo, vo, mm2)
{end *vector at termination and reflection*}

END;
{end *loop of iterations*}

(write icom on file*)
  IF wri THEN BEGIN
    assign(dout, filename+.dat'); rewrite(dout);
    writeln(dout, nop+1); {write number of points on line}
    writeln(dout, posi); {write length of differential section}
    FOR k:=0 TO nop DO BEGIN
      cadd(icom, itot[k,1], itot[k,2]);
      writeln(dout, k*posi:15, ',', icom.r, ',', icom.i, ',', ca(icom)*1e6)
    END;
    close(dout)
END;
{end *write icom on file*}
cls;
write('do you want a new configuration ? ');
ask for an other file
yesno(again,valid)
END
(end *loop for more files*)
END.
APPENDIX B: SOURCE CODE OF RAD.

PROGRAM rad(input,output,con,din,dout);
CONST pi = 3.1415926536; tpi = 6.2831853072; fpi = 12.566370614; pit = 1.5707963268; eps = 8.8419412836E-12; rt2 = 1.4142135623; lnlO = 2.302585093; maxwire = 1000;
TYPE complex = RECORD
  r : real;
  i : real
END;
comparr = ARRAY [1..maxwire] OF complex;

PROCEDURE rtoc(VAR z:complex; amp, arg:real);
  { conversion real to complex }
BEGIN
  z.r := amp * cos(arg); z.i := amp * sin(arg)
END;

FUNCTION ca(x:complex):real; { amplitude \(a\) of complex }
BEGIN
  ca := sqrt(sqr(x.r) + sqr(x.i));
END;

FUNCTION cp(x:complex):real; { phase \(\arg(x)\) of complex }
VAR h : real;
BEGIN
  WITH x DO
  BEGIN
    IF r=0 THEN h := pit
    ELSE
      h := arctan(i/r);
      IF r<0 THEN BEGIN IF i<0 THEN h := h-pi ELSE h := h+pi END
  END;
  cp := h
END;

PROCEDURE cadd(VAR z:complex; x,y:complex);
  { \(z := x + y\) \(z, x, y:\) complex }
BEGIN
  z.r := x.r + y.r; z.i := x.i + y.i
END;

PROCEDURE csub(VAR z:complex; x,y:complex);
  { \(z := x - y\) \(z, x, y:\) complex }
BEGIN
  z.r := x.r - y.r; z.i := x.i - y.i
END;
PROCEDURE cmul(VAR z:complex; x,y:complex);
{ z:=x*y  z,x,y:complex }
BEGIN
  z.r:=x.r*y.r-x.i*y.i;z.i:=x.r*y.i+x.i*y.r
END;

PROCEDURE crmul(VAR z:complex; x:complex; y:real);
{ z:=x*y  z,x:complex;y:real }
BEGIN
  z.r:=x.r*y;z.i:=x.i*y
END;

PROCEDURE cls; { clear screen }
VAR i: integer;
BEGIN
  FOR i:=1 TO 40 DO writeln
END;

PROCEDURE yesno(VAR answer:boolean); { ask for answer yes or no }
VAR yn: char;
BEGIN
  answer:=false;
  readln(yn);
  IF (yn='y') OR (yn='Y') THEN answer:=true
END;

VAR
  i,j,nofi,nofp           : integer;
  more                  : boolean;
  i0,il,hxi,hyi,exi,eyi,ezi,hxj,hyj,exj,eyj,ezj,ph1,ph2
                              : complex;
  ha1,ha2,hb1,hb2,hc1,hc2,hd1,hd2,ha,icomj
                              : complex;
  icom                  : comparr;
  inname, outname       : string;
  con,din,dout,cout     : text;
  h1,h2,hght,lngth,efr,rfr,k,ldif,r1q,r1,r2q,r2,distx
                              : real;
  a1,a2,a3,b1,b2,b3,xp,yp,zp,zd,zi,yi,y2,lenx
                              : real;
  y1xp,y2xp,kr1,kr2,kkriq,kkriq,ldif1,ldif2,kk,kkk,over
                              : real;
BEGIN
  more:=true;
{ *loop for files* }
  WHILE more DO
    BEGIN

{*read configuration*}
cls;
write('input filename = ');readln(inname);
write('output filename = ');readln(outname); writeln;
assign(con,inname+'.con');reset(con);
readln(con,over);readln(con,over);readln(con,over);
readln(con,h1);h1:=h1/1000; writeln('height conductor1 = ',h1);
readln(con,h2);h2:=h2/1000; writeln('height conductor2 = ',h2);
hght:=(h1+h2)/2; {height of straight wire configuration}
readln(con,over);
readln(con,length); writeln('length of bus = ',length);
readln(con,efr); writeln('frequency = ',efr);
close(con); {close configuration file}
rfr:=tpi*efr;k:=rfr/3E8;kk:=k*k;kkk:=k*k*k;
writeln; writeln('line of observation points ');
write('begin point');
write('b1 = ');readln(b1);
write('b2 = ');readln(b2);
write('b3 = ');readln(b3);
writeln('increment vector');
write('a1 = ');readln(a1);
write('a2 = ');readln(a2);
write('a3 = ');readln(a3);
write('number of increments = ');readln(nofi);
{end *read configuration*}

{*write observation line*}
assign(cout,outname+'.obs');rewrite(cout);
writeln(cout,inname); {name of configuration file}
writeln(cout,b1,' ',bZ,' ',b3); {begin point}
writeln(cout,a1,' ',aZ,' ',a3); {increment vector}
writeln(cout,nofi); {number of increments}
close(cout);
{end *write observation line*}

{*read common-mode current on wire*}
cls; writeln('reading data');
assign(din,inname+'.dat'); reset(din);
readln(din,nofp); {read number of points on wire}
read(din,ldif); {read length of differential section}
read(din,over);read(din,i0.r);readln(din,i0.i);
ldif1:=ldif*kk/fpi;ldif2:=ldif*kkk/(fpi*rfr*eps);
{loop for points*}
FOR i:=l TO nofp DO BEGIN
read(din,over);read(din,i1.r);readln(din,i1.i);
icom[i].r:=0.5*(i0.r+i1.r);icom[i].i:=0.5*(i0.i+i1.i);i0:=i1
END;
{end *loop for points*}
close(din); {close input file}
{end *read common-mode current on wire*}
(*calculation of radiation*)
cls; writeln('working');
assign(dout, outname + '.dts'); rewrite(dout); {open output file}
xp = bl; yp = b2; zp = b3; distx = 0; lenx = sqrt(a1 * a1 + a2 * a2 + a3 * a3);
(*loop observation line*)
FOR i := 0 TO nofi DO
BEGIN
(*radiation in point of observation*)
yl := yp - hght; y2 := yp + hght; ylxp := y1 * yl + xp * xp; y2xp := y2 * y2 + xp * xp;
rto(hxi, 0, 0); hyi := hxi; exi := hyi; eyi := exi; ezi := eyi;
{initialization}
zl := idlf / 2;
(*loop radiating wire*)
FOR j := 1 TO nofp DO
BEGIN
zd := zp - zl; icomj := icom[j];
rlq := ylxp + zd * zd; rl1 := sqrt(rlq); kr1 := k * rl1; kkr1q := k * rlq;
r2q := y2xp + zd * zd; r2 := sqrt(r2q); kr2 := k * r2; kkr2q := k * r2q;
phl.r := cos(kr1); phl.i := -sin(kr1); {phase shift}
ph2.r := cos(kr2); ph2.i := -sin(kr2);
hal.r := 1 / (kkr1q); hal.i := 1 / (kr1); cmul(ha1, hal, phl);
hb1.r := hal * y1 / rl; hb1.i := hal * y1 / rl; {Hx conductor 1}
ha2.r := 1 / (kkr2q); ha2.i := 1 / (kr2); cmul(ha2, ha2, ph2);
hb2.r := ha2 * y2 / r2; hb2.i := ha2 * y2 / r2; {Hx conductor 2}
ha := hb1 - hb2; ha.i := hb1 - hb2; cmul(ha, ha, icomj);
hxj.r := ha.r * idlf; hxj.i := ha.i * idlf;
hx1.r := hxi.r + hxj.r; hx1.i := hxi.i + hxj.i; {Hy conductor 1}
ha := hal * xp / rl; ha.i := hal * xp / rl; {Hy conductor 1}
hb2.r := ha2 * y2 / r2; hb2.i := ha2 * y2 / r2; {Hy conductor 2}
ha := hb1 - hb2; ha.i := hb1 - hb2; cmul(ha, ha, icomj);
hj.r := ha.r * idlf; hj.r := ha.r * idlf;
hj.i := hyi.r + hj.r; hj.i := hyi.i + hj.i; {Hy contribution}
hal.r := 3 / (kkr1q); hal.i := 1 - 3 / (kkr1q) / (kr1);
hal.r := hcr1.r * zd * xp / rlq; hal.i := hcr1.i * zd * xp / rlq;
{Ex conductor 1}
hb2.r := 3 / (kkr2q); hb1.i := 1 - 3 / (kkr2q) / (kr2);
cmul(hc2, hb2, ph2);
hd2.r := hcr2.r * zd * xp / rlq; hd2.i := hcr2.i * zd * xp / rlq;
{Ex conductor 2}
ha := hd1 - hd2; ha.i := hd1 - hd2; cmul(ha, ha, icomj);
exj.r := exi.r + exj.r; exj.i := exi.i + exj.i; {Ex contr.}
hd1.r := hcl.r * zd * y1 / rlq; hd1.i := hcl.i * zd * y1 / rlq;
{Ey conductor 1}
hd2.r := hcl2.r * zd * y2 / r2q; hd2.i := hcl2.i * zd * y2 / r2q;
{Ey conductor 2}
ha := hd1 - hd2; ha.i := hd1 - hd2; cmul(ha, ha, icomj);
eyj.r := eyi.r + eyj.r; eyj.i := eyi.i + eyj.i; {Ey contr.}
hal.r := 2 / (kkr1q); hal.i := -2 / (kkr1q * kr1);
hd1 := hcl.r * ylxp / rlq; hcl.i := hcl.i * ylxp / rlq; hal.i := hal.i - hcl.i;
cmul(ha1,ha1,phi1);  (Ez conductor 1)
ha2.r:=2/(kkr2q);ha2.i:=-2/(kkr2q*kr2);
hc2.r:=hb2.r*y2xp/r2q;hc2.i:=hb2.i*y2xp/r2q;
ha2.r:=ha2.r-hc2.r;ha2.i:=ha2.i-hc2.i;
cmul(ha2,ha2,phi2);  (Ez conductor 2)
ha.r:=ha1.r-ha2.r;ha.i:=ha1.i-ha2.i;cmul(ha,ha,icomj);
ezj.r:=ha.r*ldif2;ezj.i:=ha.i*ldif2;
ezi.r:=ezi.r+ezj.r;ezi.i:=ezi.i+ezj.i;  (Ez contr.)
zl:=zl+ldif  (new position on radiating wire)
END;
{end *loop radiating wire*}
{end *radiation in point of observation*}

{*save on disk*}
write(dout,distx:15,' ');  (position along observation line)
write(dout,ca(hxO-le6,' ',ca(hyO-le6,' '));
  { :Hx:, :Hy: (uA/m) }
writeln(dout,ca(exO-le6,' ',ca(eyi)-le6,' ',ca(ezi)-le6);
  { :Ex:,:Ey:,:Ez: (uV/m) }
{end *save on disk*}

write(i mod 10:1);  (wait indication)
xp:=xp+a1;yp:=yp+a2;zp:=zp+a3;  (new point of observation)
distx:=distx+lenx
END;
{end *loop observation line*}
{end *calculation of radiation*}

close(dout);  (close output file)

{*ask for more files*}
cls;
write('do you want to enter an other file ? ');
yesno(more)
{end *ask for more files*}

END
{end *loop for files*}
END.