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CALCULATING THE EFFECTIVE MECHANICAL DAMPING OF PARALLEL TURBO-GENERATORS

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G. D. Jennings

R. G. Harley

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Abstract - Torsional mode damping is a critical parameter in studies to assess the stability of a turbo-generator particularly when connected to a series capacitor compensated transmission system. In a multi-generator system similar generators are often lumped into a single equivalent generator for analysis. Although many of the generator parameters may be similar or even identical, their torsional mode dampings depend on the electrical loading of each generator and therefore often have quite different values; in such a case it is not simple to calculate an accurate value for the equivalent or effective torsional damping of the single equivalent generator. Previous researchers have proposed an *algorithm* to calculate this *effective damping*; this paper further analyses the problem of effective damping using eigenvalue methods and compares the results with those obtained using the algorithm. It shows that for certain torsional modes of a turbo-generator the algorithm may yield significant errors, and that the magnitude of these errors depend upon the torsional properties of the turbo-generator.

INTRODUCTION

The multi-inertia shafts of turbo-generators are susceptible to interaction with the electrical resonances of series capacitor compensated transmission lines and this can lead to a form of instability known as subsynchronous resonance (SSR). In SSR analyses the shaft dynamics of such generators can be represented by a number of torsional modes of oscillation [1]. The coupling (or interaction) of a mechanical torsional mode to the electrical oscillations through the electrical torque of the generator results in a combined system mode. The oscillations of this system mode become unstable if the negative contribution to the system damping (due to the electrical resonances of the transmission line) exceeds the inherent mechanical damping of the mode.

In order to avoid the destructive consequences of these unstable oscillations, SSR stability studies are necessary to establish the safe regions of system operation and to determine the extent to which any countermeasures may be employed to improve the stability limits. The two analysis techniques most often employed to assess SSR stability are eigenvalue calculations and frequency scanning methods.

The frequency scanning method provides an approximate calculation of the electrical system's contribution to the combined system mode damping. A negative electrical damping contribution which exceeds the mechanical damping of the torsional mode will then indicate that the combined system mode is unstable. The eigenvalue technique provides the nett damping of the combined system mode directly as the negative of the real part of the corresponding eigenvalue determined from linearisation of the full system equations. Although the eigenvalue technique is more accurate it has a disadvantage in that the calculations, which are computationally intensive, need to be performed repeatedly for every envisaged system condition.

In the case of two or more parallel turbo-generators feeding a series compensated transmission line, assessment of SSR stability is complicated by the interaction between the generators themselves [6-8]. The individual modes of each generator in a two machine system are coupled into a pair of system modes known as the common mode and the anti-mode, in which the

torsional oscillations of the machines occur in the same sense and in the opposite sense respectively. In order to assess the stability of the system the two machines are often lumped into a single *equivalent generator* with base MVA rating equal to the combined MVA rating of the two machines. However this is only a valid simplification for truly identical equally loaded machines, where the effective mechanical damping of the single *equivalent generator* (presented by the coupling of the generators in the common mode) is the same as the torsional mechanical damping of either of the two machines, since they are identical.

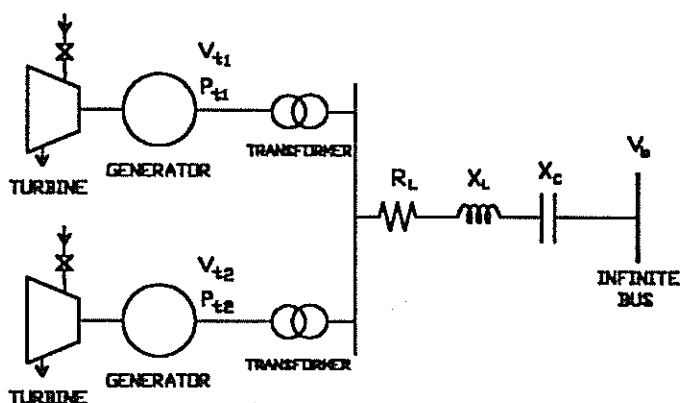


Fig. 1 Two generator system under study

However, if the two generators are operating at different levels of electrical loading, they will have different torsional mechanical dampings. In such cases the effective mechanical damping presented by the coupling of the two machines in the common mode is not equal to the mechanical damping of either of the two generators, but lies somewhere between these values. Thus it is apparent that when unequally loaded generators are synchronised in parallel to a common transmission line as in Fig. 1 there is some sharing of damping between the generators in the common mode.

Unfortunately for unequally loaded generators it is not possible to obtain an explicit expression which describes how the individual generator mechanical dampings combine to make up the effective mechanical damping of the coupled common mode oscillations. Thus it is not always possible to assign a representative mechanical damping value to a single equivalent generator such that it can represent a group of generators for SSR analysis. This creates a problem in the use of the frequency scanning technique since this method calculates purely the negative damping contribution of the electrical system to the total system mode damping, which must then be added to the effective mechanical damping of a *single equivalent generator* (but for which an explicit accurate expression does not exist).

Eigenvalue calculation provides numerical values of the nett damping of the combined two machine system modes (common and anti-modes) directly for any combination of individual mechanical dampings of the two machines. This enables the stability of the system to be assessed without the need to know the explicit relationship between the generator mechanical dampings and the common mode dampings. However repeated eigenvalue calculations are costly and time consuming for large

multi-machine systems. A closed form expression to approximate the effective damping presented by two or more unequally loaded turbo-generators is therefore desirable so that this function could be used to describe the mechanical damping of a single equivalent generator for more rapid stability assessment. It is important that the validity of such an expression be thoroughly established, since torsional damping is a critical parameter affecting the stability of turbo-generators connected to series compensated transmission systems. Furthermore, an understanding of the way in which unequally loaded generators share torsional damping would provide a useful tool for assessing the SSR countermeasure requirements in systems where the proportion of the load carried by each generator varies considerably.

A recent study by Agrawal and Farmer [3] proposed an approximate algorithm to describe the effective mechanical damping of unequally loaded identical turbo-generators. They show that their algorithm can provide a mechanical damping value for an equivalent single generator which accurately represents the behaviour of two unequally loaded generators. This paper revisits the assumptions of Ref. [3] and presents additional theoretical results to show that the algorithm proposed in Ref. [3] can lead to significant errors for different combinations of system parameters.

The system configuration under investigation is the same as that in Ref. [3] and consists of two generators synchronised in parallel through a series compensated transmission line to an infinite bus as shown in Fig. 1. The generator and transmission line parameters used for the theoretical studies in the paper are those of a laboratory micro-alternator system at the University of Natal. These micro-alternators were designed such that their characteristics closely match those of two large turbo-generators [9] on a per-unit basis, and hence the system studied is representative of an actual system.

THE EFFECTIVE DAMPING ALGORITHM [3]

The authors of Ref. [3] propose that the effective damping σ_{ef} of two turbo-generators in parallel can be related analytically to the individual mechanical dampings of each generator by the following algorithm (also called the *effective damping algorithm* or EDA for the purpose of this paper)

$$\frac{1}{\sigma_{ef}} = \frac{1}{2} \left[\frac{1}{\sigma_1} + \frac{1}{\sigma_2} \right] \quad (1)$$

where σ_1 and σ_2 are the individual mechanical dampings of generators 1 and 2 (in isolation from the electrical system) described in their unique form as mode decrement factors. Thus according to the EDA the damping in parallel generators is combined into the system's common mode damping in a similar manner to resistors in parallel. The most important of the assumptions made by Ref. [3] in arriving at the expression of eqn. (1) are as follows:

- (i) The generators are identical in all respects apart from the differences in their mechanical mode dampings.
- (ii) The generators are assumed to be at incipient stability.
- (iii) Under incipient stability conditions the oscillations of both generators in the mode of interest can be assumed to be in phase.

The incipient stability condition of a particular torsional mode occurs when the undamping due to the electrical network is exactly equal to the positive mechanical damping of the torsional mode, resulting in its nett damping in the total system being zero. For the two machine system this means that the undamping due to the electrical network is exactly equal to the effective damping σ_{ef} of the machines in parallel.

The authors of Ref. [3] studied a tuned radial transmission system fed by two generators such as that in Fig. 1, with high undamping due to the electrical network. The frequency

scanning method was used to calculate the maximum undamping $\Delta\sigma$ due to the electrical network for the mode being studied. Several combinations of generator mechanical dampings σ_1 and σ_2 were chosen such that the effective damping values σ_{ef} predicted by the algorithm of eqn. (1) were close to the undamping due to the electrical network (thus ensuring incipient stability conditions).

In order to verify their EDA the authors [3] then performed an eigenvalue study for each of these damping combinations to determine the *nett damping* of the torsional common mode (negative of eigenvalue real part). By adding the known undamping $\Delta\sigma$ due to the electrical network to this *nett damping* in each case, the effective damping was thus also arrived at through eigenvalue analyses. The effective dampings predicted by the EDA of eqn. (1) for each combination of generator dampings showed good agreement with the effective dampings determined in this way from the common mode eigenvalues.

EFFECTIVE DAMPING RESULTS FOR STUDY SYSTEM

Several eigenvalue analyses are carried out in this paper on a linearised model of the two generator system shown in Fig. 1 in order to investigate the accuracy of applying the EDA to the laboratory micro-alternator system. Details of the various models used to describe the system can be found in Refs. [4] and [5]. The approach adopted to study the effective damping of a particular mode in each study system is outlined below:

An eigenvalue scan is performed to determine the amount of series capacitance required in the transmission line of the micro-alternator system in order that the nett torsional damping of the particular mode in the total system (the common mode decrement factor σ_c) is at its most unstable point. The mechanical dampings σ_1 and σ_2 of both generators in the computer program are then increased equally until the common mode damping is zero, thus bringing the mode to incipient stability. This serves as a *base case* for the study of this mode since the effective damping in the case of identically damped generators can be determined analytically and is known to be equal to the damping of either generator. In the base case the decrement factor σ_c of the system mode for this value of series capacitance is zero and hence the undamping $\Delta\sigma$ due to the electrical network is known to be equal to the mechanical damping of each generator. By extension, the undamping due to the electrical network is the same for any combination of dampings of the generators if the value of the series capacitance is kept constant.

The mechanical mode damping σ_1 of the first generator is then increased in the computer program, and the EDA [eqn. (1)] is used to determine the mechanical mode damping of the second generator σ_2 that results in the same effective damping as is presented by the identically damped generators in the base case. The series capacitance is held fixed at its base case value so that the undamping due to the electrical network remains fixed throughout. Thus if the effective damping predicted using the EDA is correct, the common mode decrement factor of the system at the new damping condition, determined from the corresponding eigenvalue, should remain zero.

The effective damping for each set of individual generator dampings is obtained through eigenvalue analysis by adding the undamping $\Delta\sigma$ determined from the base case, to the negative of the eigenvalue real part (common mode decrement factor σ_c) as shown in Note (3) below. This is then compared to the effective damping predicted by the EDA. The difference between these two values is expressed as a percentage of the value determined from eigenvalue analysis to find the error in the EDA in each case [Note (4)].

The behaviour of the system of Fig. 1 is first predicted assuming the parameters of the system to be those of the laboratory micro-alternators. The characteristics of the multi-inertia shafts

of these machines were designed to closely match those of an actual large system on a per-unit basis, and the transmission line parameters used in this study have been chosen to correspond to the same large system.

Tables 1 to 3 show the effective damping values arrived at through eigenvalue analysis, compared with those predicted by the EDA for the first three torsional modes of this system under four different conditions of individual generator dampings.

Throughout the analyses in Table 1 the single fixed value of series capacitance is chosen such that the nett torsional damping of the combined system in mode 1 is at its most unstable value. The first row of Table 1 shows the details of the base case for the study of this mode. Here the mechanical dampings σ_1 and σ_2 in the computer program are set to be equal and their magnitude increased to 2.79 s^{-1} in order that the nett damping of the common mode σ_c in the combined system is zero for this least stable value of series compensation (incipient stability). The undamping value $\Delta\sigma$ is thus known to be equal to 2.79 s^{-1} for the fixed value of compensation used in the rest of the analyses in Table 1. For this identical damping condition in the two generators the EDA of eqn. (1) yields an effective damping value of 2.79 s^{-1} . Addition of the known undamping value $\Delta\sigma = 2.79 \text{ s}^{-1}$ to the common mode decrement factor yields the same value of effective damping from eigenvalue analysis, and hence the error in the EDA is zero for the base case.

In the second row of Table 1 the value of the mechanical damping of generator 1 has been increased to 1.2 times its previous base case value. However, the mechanical damping value of generator 2 has been decreased such that the effective damping predicted by the EDA is unchanged from the base case (2.79 s^{-1}). If the actual effective damping is unchanged the decrement factor of the common mode σ_c should remain zero, but eigenvalue analysis shows that its value is now 0.015 s^{-1} . Thus the eigenvalue method indicates that the effective damping is $0.015 \text{ s}^{-1} + \Delta\sigma = 2.805 \text{ s}^{-1}$, and that there is an error of $100 \times (2.805 - 2.790) / 2.805 = 0.6\%$ in the EDA in this case. The remaining rows of Table 1 show the same comparison for three more combinations of different generator mechanical damping values for which the EDA predicts the effective damping to be unchanged from the base case.

Tables 2 and 3 extend the above analyses to modes 2 and 3 of the micro-alternator system respectively. Although the undamping $\Delta\sigma$ of the three modes is different, the values of σ_1 and σ_2 for cases 1,2,3 and 4 in each Table are chosen to be the same multiples of the generator damping in the base case, to allow direct comparison between the modes.

TABLE 1

Effective damping - Micro-alternator mode 1 - (5.93 Hz)
($\Delta\sigma = 2.79 \text{ s}^{-1}$)

	σ_1	σ_2	σ_c (1)	σ_{ef} [EDA] (2)	σ_{ef} [Eval.] (3)	Error (4)
BASE CASE	2.790	2.790	0.000	2.790	2.790	0.0%
CASE 1	3.350	2.390	0.015	2.790	2.805	0.6%
CASE 2	4.190	2.090	0.056	2.790	2.846	2.0%
CASE 3	5.580	1.860	0.108	2.790	2.898	3.8%
CASE 4	8.370	1.674	0.149	2.790	2.939	5.3%

Notes: (1) Common mode dec. factor from eigenvalue study

(2) Effective damping from EDA of eqn. (1)

(3) σ_{ef} [Eval.] = $\sigma_c + \Delta\sigma$

(4) Error = $\frac{\sigma_{ef}[\text{Eval.}] - \sigma_{ef}[\text{EDA}]}{\sigma_{ef}[\text{Eval.}]} \times 100\%$

TABLE 2

Effective damping - Micro-alternator mode 2 - (10.80 Hz)
($\Delta\sigma = 0.298 \text{ s}^{-1}$)

	σ_1	σ_2	σ_c (1)	σ_{ef} [EDA] (2)	σ_{ef} [Eval.] (3)	Error (4)
BASE CASE	0.298	0.298	0.000	0.298	0.298	0.0%
CASE 1	0.358	0.255	0.004	0.298	0.302	1.3%
CASE 2	0.447	0.224	0.016	0.298	0.314	5.4%
CASE 3	0.596	0.199	0.034	0.298	0.332	11.4%
CASE 4	0.894	0.179	0.049	0.298	0.347	15.1%

TABLE 3

Effective damping - Micro-alternator mode 3 - (13.80 Hz)
($\Delta\sigma = 0.150 \text{ s}^{-1}$)

	σ_1	σ_2	σ_c (1)	σ_{ef} [EDA] (2)	σ_{ef} [Eval.] (3)	Error (4)
BASE CASE	0.150	0.150	0.000	0.150	0.150	0.0%
CASE 1	0.180	0.129	0.003	0.150	0.153	2.0%
CASE 2	0.225	0.113	0.011	0.150	0.161	7.3%
CASE 3	0.300	0.100	0.022	0.150	0.172	14.7%
CASE 4	0.450	0.090	0.032	0.150	0.182	21.3%

The previous paragraphs have outlined how the eigenvalue method and the algorithm proposed in Ref. [3] have been used to calculate the effective mechanical damping presented by the various modes of the two micro-alternators when they are operating in parallel. Before considering the results it worthwhile briefly summarising the approach that has been adopted, and pointing out the significance of the results shown in the tables in terms of generator stability. In each table, the series capacitance has been chosen such that the electrical undamping due to the transmission line is a maximum for the mode in question. The base case values of mechanical damping for the micro-alternators are then chosen so that the nett damping of this mode in the combined system is zero. This approach is necessary since the algorithm of Ref. [3] is valid only for this incipient stability condition. Thus in terms of the stability of the machines, the EDA predicts the combinations of individual generator mechanical dampings that are necessary for a mode of the two generator system to be stable for all compensation levels.

The results of Tables 1 to 3 show that for each mode the error in the effective damping predicted by the EDA increases from case 1 to case 4, as the difference between the mechanical damping values σ_1 and σ_2 of the two generators increases. The loss in accuracy is most pronounced in mode 3 (Table 3) where the error in the effective damping predicted by the EDA increases from 2.0% in case 1 to 21.3% in case 4, whilst in mode 2 (Table 2) the error increases from 1.3% in case 1 to 15.1% in case 4. The effective dampings predicted by the EDA are clearly most accurate for mode 1 of this particular system, where the error for the largest damping difference (case 4) is 5.3%.

The accuracies of the EDA for mode 1 of this particular system are similar to the values reported in Ref. [3], but for modes 2 and 3 of this system the EDA shows significant errors. Moreover, the errors in the effective damping values of the mode in Ref. [3] did not increase with increasing differences in generator dampings.

The loss in accuracy of the EDA from case 1 to case 4 increases from mode 1 to mode 3 of the micro-alternator system and

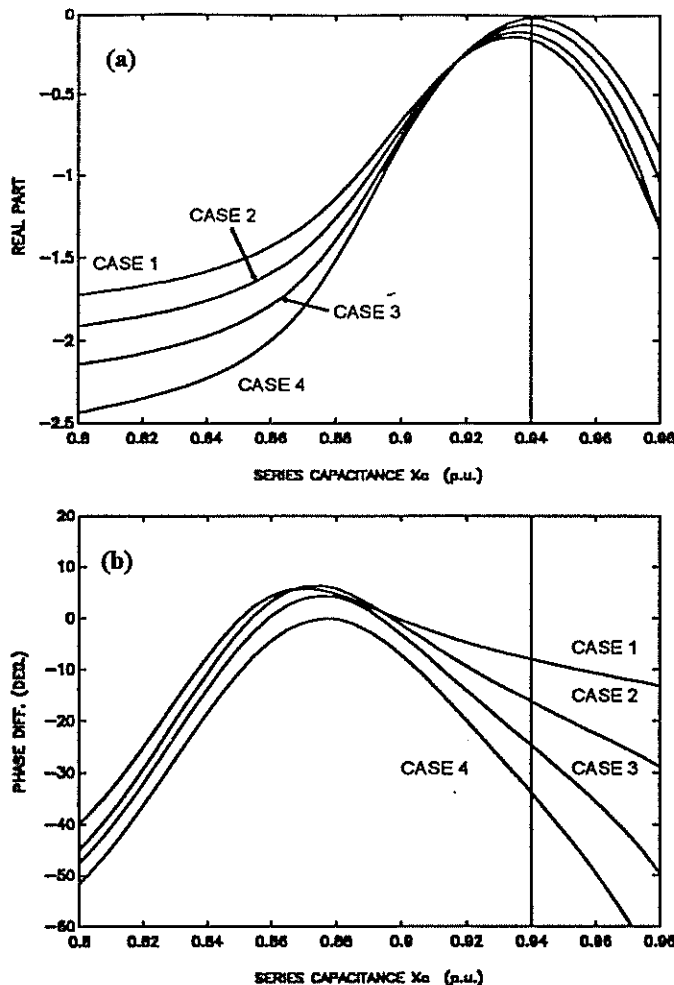


Fig. 2 - Real part of eigenvalue and phase difference between generators near incipient stability of mode 1 of the micro-alternator system.

coincides with the increasing torsional natural frequency of the modes, as well as a reduced undamping of these modes by the electrical system. This may indicate that the accuracy of the effective damping predicted for a mode is dependent on its torsional characteristics.

The results reported here for the micro-alternator system suggest that the EDA would be sufficiently accurate for determining the amount of countermeasure required to stabilise a mode with large undamping due to the electrical system such as mode 1. However the mechanical damping values of the micro-alternators which are required to stabilise mode 1 are considerably larger than the damping values typical of large turbo-generators. The potential for turbo-generator modes to become undamped to this extent in series compensated systems is not uncommon and such a problem is known to exist [9] for the turbo-generators upon which the micro-alternator system is based. However the accuracy of the EDA for a mode such as mode 1 of these machines is of little benefit since it is not practically possible to provide the additional damping required to stabilise such a mode. Rather, compensation levels which result in destabilisation of this mode of the generators are avoided in practice. However the results also suggest that the EDA would be less accurate if used to assess whether a stability problem exists for torsional modes of the system which are near incipient stability for realistic values of mechanical damping such as modes 2 and 3.

INVESTIGATION OF THE ASSUMPTIONS MADE IN DERIVING THE EFFECTIVE DAMPING ALGORITHM

In order to determine whether the assumptions (made by Ref. [3]) underlying the EDA are valid for the system studied in this paper, mode 1 of the system is subjected to a more detailed eigenvalue analysis. The first of the assumptions is that the generators are identical in all respects apart from the differences in their mechanical mode dampings. This condition has been met in both study systems considered here and thus the validity of this assumption has been ensured.

The remaining two assumptions state that the generators must be at incipient stability, and that under these conditions the generator oscillations can be assumed to be in phase. In order to test this assumption it is necessary to determine the relative phase shifts of the two generators in the mode shape vector of the particular mode as determined by the corresponding system eigenvector.

Fig. 2(a) shows the real part of the common mode eigenvalue of mode 1 of the micro-alternator system as the series compensation varies near the condition of incipient stability. The four curves correspond to the four damping cases considered for this mode in Table 1. Fig. 2(b) shows the corresponding variation in the phase difference between the generator rotor speeds in mode 1 for the four damping cases. The actual value of series capacitance adopted in Table 1 for this mode is 0.94 p.u. Fig. 2(a) clearly shows that at this compensation level the mode 1 eigenvalue for each damping case is near incipient stability. However Fig. 2(b) shows that at this incipient stability condition there is an increasing phase shift between the generators from case 1 to case 4. Recalling the results of Table 1, this corresponds to an increase in the error of the effective damping predicted by the EDA for mode 1.

Fig. 2(b) shows that for the system studied in this paper the EDA assumption that the oscillations of two generators with different mechanical dampings are in phase under incipient stability conditions, is not valid for mode 1. Furthermore, the assumption becomes increasingly more invalid as the differences between the mechanical dampings of the generators increase. Similar studies confirm (although not shown here) that this trend occurs for modes 2 and 3 as well.

The question that then arises is whether the error observed in the effective dampings predicted by the EDA (which was derived assuming no phase shift between the generator rotors) can be directly related to the actual magnitude of the phase shifts encountered. Based on the results shown here for mode 1 (Fig. 2) and the results of similar analyses for modes 2 and 3 of the micro-alternator system, it clearly holds that for any one mode of the two systems the increasing error in the EDA is accompanied by an increase in the phase difference between the generator rotor speeds. However for a given magnitude of phase shift between the generator rotors the error in the EDA is markedly different for two torsional modes of the same system.

The results of Tables 1 to 3 indicate the possibility that the error in the EDA is related to the torsional properties of the generator shafts. The subsequent investigation into the assumptions made in deriving the EDA show that in fact the oscillations of the generator rotors can not be assumed to be in phase at incipient stability for any of the modes considered in these studies. This is true even in the case of mode 1 of the micro-alternator system for which the EDA is found to be sufficiently accurate. Thus the subsequent investigation into the EDA assumptions has been unable to demonstrate that the accuracy of the EDA in Tables 1 to 3 goes hand in hand with the validity of a particular assumption. Nonetheless, it has shown that one assumption is not valid for any of the modes studied and that in certain modes this is accompanied by significant errors in the EDA.

It is worth noting that in all the studies reported in this paper the error in the EDA is conservative; the effective damping predicted by the EDA, although in error, is smaller than the value determined from eigenvalue analysis. Thus the system may be more stable than would be indicated by an assessment of system stability using the EDA and consequently the error incurred by its use would result in failure to safety.

CONCLUSIONS

Ref. [3] describes a closed form algorithm to calculate the effective damping to be applied to a single equivalent generator in order that such a generator may be used in SSR studies to represent the mechanical damping presented by unequally loaded parallel identical turbo-generators. This paper has reconsidered the application of this algorithm to a further study systems and investigated the validity of the assumptions upon which the algorithm is based. The results indicate that:

- In all the torsional modes of the system studied the error in the effective damping predicted by the algorithm increases as the difference between the mechanical dampings of the generators increases. For some torsional modes the error can be as large as 20% for differences in mechanical damping which can realistically be expected in large unequally loaded turbo-generators.
- The assumption made in Ref. [3] that the oscillations of generators with different mechanical mode dampings can be assumed to be in phase is not valid for the torsional modes studied in this paper. In all the modes considered in this paper the phase difference between the generator rotor speeds at incipient stability increases as the difference in the mechanical dampings of the generators increases.
- The degree of impact of this increasing phase difference between the generator rotor speeds on the accuracy of the effective damping algorithm depends on the particular torsional mode being considered. In some modes studied large increases in the phase difference between the generator rotor speeds are accompanied by significant increases in the error of the effective damping predicted by the algorithm. In other modes studied similar large increases in the phase difference between the generator rotor speeds are accompanied by only small increases in the error of the effective damping algorithm. This suggests that the torsional properties of a particular mode may be important in determining the impact which the phase difference between the generators has on the accuracy of the algorithm. In particular the results indicate that the algorithm is least accurate for higher frequency torsional modes with low undamping due to the electrical system.

The value of the effective damping algorithm lies in the considerable reduction of the computational overhead necessary to assess turbo-generator stability as opposed to the more rigorous eigenvalue methods which it has been compared to in this paper. The analyses presented here have shown that the high accuracy of the algorithm reported in Ref. [3] is valid only for modes which show large undamping due to the electrical system, where such accuracy is of little benefit. Although the results presented here indicate that the algorithm can result in considerable error for modes which are likely to be marginally stable under normal levels of turbo-generator loading, this error is conservative and thus need not necessarily discourage its use. However the results presented in the paper contribute towards a better understanding of the use of the effective damping algorithm by drawing attention to the possible inaccuracies and showing that they depend upon the properties of the torsional mode being considered.

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APPENDIX

Transmission Line Parameters

$$R_{T1} = R_{T2} = 0.001 \text{ p.u.}$$

$$X_{T1} = X_{T2} = 0.15 \text{ p.u.}$$

$$R_L = 0.0845 \text{ p.u. } X_L = 1.0097 \text{ p.u.}$$

Micro-Alternator Electrical Parameters

$$R_a = 0.006 \text{ p.u.}$$

$$X_a = 0.11 \text{ p.u.}$$

$$X_{md} = 1.98 \text{ p.u.}$$

$$X_{mq} = 1.87 \text{ p.u.}$$

$$X_f = 0.1 \text{ p.u.}$$

$$R_f = 0.001 \text{ p.u.}$$

$$X_{kd} = 0.125 \text{ p.u.}$$

$$R_{kd} = 0.0212 \text{ p.u.}$$

$$X_{kq} = 0.257 \text{ p.u.}$$

$$R_{kq} = 0.027 \text{ p.u.}$$

Micro-Alternator Mechanical Parameters

[Q] MATRIX

Mode0	Mode1	Mode2	Mode3	Mode4	Mode5	
1.0000	-1.0000	1.0000	-1.0000	0.4245	0.0000	HP
1.0000	-0.6926	-0.0263	0.6688	-0.6177	-0.0000	LP1
1.0000	-0.2753	-0.4203	-0.1433	1.0000	0.0000	LP2
1.0000	0.2998	-0.2169	-0.6729	-0.8339	-0.0002	LP3
1.0000	0.6798	0.2502	0.3305	0.2405	0.0208	GEN
1.0000	0.6828	0.2538	0.3384	0.2491	-1.0000	EXC
$J_{Mi} \text{ (p.u.)}$						
0.0362	0.0123	0.00474	0.0106	0.0169	0.000224	
$v_i \text{ (Hz)}$						
0.00	5.93	10.80	13.80	16.80	91.20	
Torque dist. (%)						
33.7	21.6	18.8	25.9	0.0	0.0	

ALGORITHM FOR THE CALCULATION OF TRANSIENT PROCESSES IN LARGE ELECTRICAL POWER SYSTEMS

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1. Abstract.

The report considers the ways of developing flexible and reliable algorithm of calculation of electrical power system transients under large disturbances. For this purpose it is proposed to use convolution integral for the calculation of linear dynamic elements responses. This refers first of all to automatic control systems. Such an approach reduces both the order of differential equations system and its "stiffness". Besides, the set of equations to be solved at each integration step is maximally reduced to that of load flow, allowing for the wide use of well known algorithmic and programming developments.

2. Introduction.

Computations of electromechanical transients in electrical power systems (EPS) caused by large disturbances are extensively used in planning and operational practice. The aim of these computations is to verify the transient stability of the system and to estimate the effect of various automatic control and protection systems.

EPS growing in size and complexity and degree of their automation constantly increasing, the results of the transients computations become increasingly important. Sufficient to say that malfunction or maladjustment of control or protection device may result in practically unpredictable development of the transients.

Correct description of complex EPS transients with due account of automatic devices results in a system of differential and algebraic equations of a very high order, and besides the differential equations subsystem is "stiff" as the values of time constants of regulators are usually much smaller than those of generators. All this makes the problem of numerical solution of these equations rather critical from the viewpoint of computational efficiency and reliability of results.

The aim of this report is to outline an approach which overcomes these difficulties by the maximum use of well developed technique

of load-flow calculations and judicious application of convolution integral.

3. Mathematical Model.

The model of EPS consists of two main subsystems [1]:

1) dynamic subsystem, comprising differential equations of transients in power stations generating units;

2) static subsystem, comprising nodal algebraic equations of electrical network operating condition.

As can be seen later this restricted model can be easily generalized to include dynamic loads, SVC's and other controlled elements of EPS.

3.1 Dynamic subsystem.

This subsystem is based on Park's equations of a synchronous machine. Assuming that the machine has two damper windings we get well known 5-th order model in the form:

$$d\delta/dt = s \quad (1)$$

$$\frac{ds}{dt} = \left[T_t - E_q' i_q - E_d' i_d - i_d i_q (x_d'' - x_q'') \right] \frac{\omega_{nom}}{M} \quad (2)$$

$$T_{d_0} \frac{dE_q'}{dt} = E_{qe} - E_q' + i_d (x_d - x_d') + [E_q'' - E_q' - i_d (x_d' - x_d'')] k_2 \quad (3)$$

$$T_{d_0} \frac{dE_d'}{dt} = [E_{qe} - E_q' + i_d (x_d - x_d')] k_1 + [E_q' - E_q'' + i_d (x_d' - x_d'')] k_3 \quad (4)$$

$$T_{q_0} \frac{dE_d'}{dt} = -E_d' - i_q (x_q - x_q'') \quad (5)$$

where

$$k_1 = [x_{ad} - (x_d - x_d'')] y T_{d_0}' / T_{d_0}, \quad k_2 = (x_d - x_d') y, \\ k_3 = 1 - k_1 k_2, \quad y = [x_{ad} - (x_d - x_d'')]^{-1}, \quad \text{the meaning of other symbols is well known.}$$

The set of differential equations (1)-(5)

is stiff as $T_{q_0}' \ll T_{q_0}$ and $T_{d_0}' \ll T_{d_0}$. Taking into account dynamic properties of automatic excitation regulator and turbine governor leads

to a substantial increase both in the order and stiffness of the differential equations resulting in significant decrease of computational efficiency.

As the regulators are represented by linear dynamic blocks and a limited number of nonlinearities, usually at the output, they can be handled using convolution integral. The same applies to the equations of damper windings (4), (5).

Consider this approach taking (5) as an example. Let all the variables be known at $t=t_n$, and it is required to determine E_d at $t>t_n$. Using stator windings equations:

$$E_d - x_q i_q = u_d, \quad E_q + x_d i_d = u_q \quad (6)$$

and substituting i_q from (6) into (5) we get the latter in the form which is suitable for use together with nodal network equations:

$$T_q \frac{dE_d}{dt} + E_d = (1 - x_q/x_q) u_d$$

Written in deviations of the variables from their values at $t=t_n$ this equation becomes:

$$T_q \frac{d\Delta E_d}{dt} + \Delta E_d = (1 - x_q/x_q) \Delta u_d - \Delta E_d + (1 - x_q/x_q) u_{d_n}$$

where: $\Delta E_d = E_d - E_{d_n}$, $\Delta u_d = u_d - u_{d_n}$.

Using Carson's transform we get:

$$\Delta E_d(p) (1 + pT_q) = \{ (1 - x_q/x_q) \Delta u_d(p) - [E_{d_n} - (1 - x_q/x_q) u_{d_n}] \}$$

Considering this equation as the description of linear first order element with time constant T_q , two inputs and one output,

one can calculate the output signal $\Delta E_d(t)$ as a response to an input signal $\Delta u_d(t)$. This response can be calculated analytically provided the input signal is conveniently approximated.

The analysis has shown that in case of well known stepwise approximation of input signal the response is very sensitive to sampling period especially for differentiating elements which are characteristic of forced excitation control [2]. For this reason more accurate - piecewise linear and quadratic approximation were tried which proved much less sensitive to the sampling period. Of these two piecewise linear approximation appeared to be quite suitable as it combines simplicity of calculations and reliability of the results [2]. Necessary analytical derivations are outlined in Appendix. Using these formulas we get:

$$E_d = E_{d_n} C_q + (1 - x_q/x_q) \times \left[(u_d - u_{d_n} C_q) - \frac{(1 - C_q)(u_d - u_{d_n}) T_q}{(t - t_n)} \right] \quad (7)$$

where $C_q = \exp[-(t - t_n)/T_q]$.

In a similar way from (3), (4) and (6) we get

$$E_q = E_{q_n} C_d + (1 - x_d/x_d') \times \left[(u_q - u_{q_n} C_d) - \frac{(1 - C_d)(u_q - u_{q_n}) T_d}{(t - t_n)} \right] + \left[(E_q' - E_{q_n} C_d) - \frac{(1 - C_d)(E_q' - E_{q_n}) k T_{d_0}}{(t - t_n)} \right] C \quad (9)$$

where $C = x_d'/x_d$, $k = (x_d - x_{ad})y(1 - x_d'/x_d)$, $C_d = \exp[-(t - t_n)/T_d]$.

Excluding i_d and i_q from equations (2) and (3) with the help of (6) and substituting into resulting equations expressions (7) and (8)

for E_d and E_q we finally get third order model for synchronous generator transients, which is not stiff, as damper windings transients are accounted for in right hand sides of differential equations:

$$d\delta/dt = s$$

$$\frac{ds}{dt} = \left[T_t - T_{d_0} (E_q', u_q, u_d, E_{q_n}, E_{q_n}', u_{q_n}, u_{d_n}, t) \right] \times \frac{\omega_{nom}}{M} \times \frac{E_{q_n} - E(E_q', u_q, E_{q_n}, E_{q_n}', u_{q_n}, t)}{T_{d_0}} \quad (9)$$

These equations can be easily integrated by any known numerical method. In a particular case of constant integration step ($\Delta t = \text{const}$) equations (7) and (8) become linear time invariant ones, as $t = t_n + \Delta t$.

The approach outlined above allows taking into account dynamic characteristics of turbine and excitation system together with corresponding automatic regulators without practically any limitations[1]. For this aim one should consider T_t and E_{q_n} as responses of dynamic elements to corresponding input signals, approximated as outlined in Appendix. The structure of mathematical model (9) will not be changed.

3.2. Static subsystem.

It is based on load flow equations, describing operating state of electrical network

together with static loads. For our purpose load flow equations in the form of nodal current balances in rectangular frame are most suitable. In this case nodal voltages and currents are represented as: $\dot{U}_i = U_i' + jU_i''$, $\dot{I}_i = I_i' + jI_i''$.

In order to combine load flow equations with those of synchronous generators one should use the following relations, linking corresponding co-ordinates in the form:

$$\begin{aligned} I_{qi}' &= i_{qi} \cos \delta_i - i_{di} \sin \delta_i, \\ I_{di}' &= i_{di} \cos \delta_i + i_{qi} \sin \delta_i \end{aligned} \quad (10)$$

$$\begin{aligned} u_{qi} &= U_i' \cos \delta_i + U_i'' \sin \delta_i, \\ u_{di} &= U_i'' \cos \delta_i - U_i' \sin \delta_i \end{aligned} \quad (11)$$

Using (6)-(8) and (11) we can rewrite (10) in the form:

$$\begin{aligned} I_{qi}' &= I_{qi}'(E_{qi}', U_i', U_i'', \delta_i, t) \\ I_{di}' &= I_{di}'(E_{qi}', U_i', U_i'', \delta_i, t) \end{aligned} \quad (12)$$

Correspondingly with the account of (11), the equations of transients in synchronous generators can be presented as:

$$\begin{aligned} ds_i/dt &= s_i \\ \frac{ds_i}{dt} &= [T_{ui} - T_{cli}(E_{qi}', U_i', U_i'', \delta_i, t)] \times \frac{\omega_{nom}}{M} \quad (13) \\ \frac{dE_{qi}'}{dt} &= \frac{E_{qei} - E_i(E_{qi}', U_i', U_i'', \delta_i, t)}{T_{dqi}} \end{aligned}$$

3.3. The mathematical model of complex EPS.

EPS model for the calculation of electromechanical transients consists of:

-- the set of differential equations, describing the behavior of controlled synchronous generators, i.e. equations (13) for each machine, which can be presented as

$$dx/dt = f(x, y, t) \quad (14)$$

-- the set of algebraic load flow equations together with equations (12) for generating nodes

$$F(x, y, t) = 0, \quad (15)$$

where x is the column of state variables for generators,

$$x = (\delta_1 s_1 E_{q1}', \delta_2 s_2 E_{q2}', \dots),$$

y is the column of nodal voltages components,
 $y = (U_1' U_1'', U_2' U_2'', \dots),$

f, F are nonlinear vector-functions.

4. The Algorithm of Calculations of Transient Processes in EPS.

The simplest implicit method of numerical integration - trapezoidal rule - has been used for the solution of equations (14) and (15). According to this method the values

of variables x and y at time instant $t_{n+1} = t_n + \Delta t$ (where Δt is integration step) are determined by the solution of the following set of algebraic equations:

$$x_{n+1} - x_n -$$

$$- [f(x_{n+1}, y_{n+1}, t_{n+1}) + f(x_n, y_n, t_n)] \Delta t / 2 = 0 \quad (16)$$

$$F(x_{n+1}, y_{n+1}, t_{n+1}) = 0 \quad (17)$$

which can be performed by Newton's method.

The advantages of such an approach are as follows:

1) the method is computationally stable and self-starting [3];

2) as equations (17) are load flow ones, one may use to maximum extent all the algorithmic and programming results developed for the load flow calculations by Newton's method, including sparsity oriented programming.

Owing to diagonal structure of equations (16) the calculations can be easily reduced to those of (17). For this purpose the following approach is used. Equations (13) are presented in the form of (16) and $s_{i(n+1)}$ is excluded with the help of the first equation of thus obtained set resulting in:

$$\begin{aligned} \alpha_i \delta_{i(n+1)} - T_{ui(n+1)} + T_{cli(n+1)} - F_{li(n)} &= 0, \\ \beta_i E_{qi(n+1)} - E_{qei(n+1)} + E_{i(n+1)} - F_{2i(n)} &= 0, \end{aligned} \quad (18)$$

where: $\alpha_i = 4M_i / (\Delta t^2 \omega_{nom})$, $\beta_i = 2T_{dqi} / \Delta t$,

$$F_{li(n)} = \alpha_i (\delta_{i(n)} + \Delta t s_{i(n)}) + T_{ui(n)} - T_{cli(n)},$$

$$F_{2i(n)} = \beta_i E_{qi(n)} + E_{qei(n)} - E_{i(n)},$$

indices in brackets correspond to the initial ($t=t_n$) and final ($t=t_{n+1}=t_n+\Delta t$) points of the integration step.

Algebraic equations (18) together with load flow equations (17) are solved by Newton's method. In this case, according to (18) we get for each generator:

$$\begin{aligned} &(\alpha_i + \partial T_{cli} / \partial \delta_i) \Delta \delta_i^{(k+1)} + (\partial T_{cli} / \partial E_{qi}') \Delta E_{qi}'^{(k+1)} + \\ &+ (\partial T_{cli} / \partial U_i') \Delta U_i'^{(k+1)} + (\partial T_{cli} / \partial U_i'') \Delta U_i''^{(k+1)} = \\ &= F_{li(n)} + T_{ui(n+1)}^{(k)} - T_{cli(n+1)}^{(k)} - \alpha_i \delta_{i(n+1)}^{(k)}, \\ &(\partial E / \partial \delta_i) \Delta \delta_i^{(k+1)} + (\beta_i + \partial E / \partial E_{qi}') \Delta E_{qi}'^{(k+1)} + \\ &+ (\partial E_i / \partial U_i') \Delta U_i'^{(k+1)} + (\partial E_i / \partial U_i'') \Delta U_i''^{(k+1)} = \\ &= F_{2i(n)} + E_{qei(n+1)}^{(k)} - E_{i(n+1)}^{(k)} - \beta_i E_{qi(n+1)}^{(k)}, \end{aligned}$$

where k is the number of iteration. It is assumed that $T_{ui} = \text{const}$, $E_{qei} = \text{const}$. Taking into account automatic regulators complicates calculations, but does not introduce any changes in the algorithm. The last two

equations are solved for $\Delta\delta_i^{(k+1)}$ and $\Delta E_{qi}^{(k+1)}$, resulting in

$$\Delta\delta_i^{(k+1)} = a_{\delta_i} + a'_{\delta_i} U_i^{(k+1)} + a''_{\delta_i} U_i^{(k+1)},$$

$$\Delta E_{qi}^{(k+1)} = a_{E_i} + a'_{E_i} U_i^{(k+1)} + a''_{E_i} U_i^{(k+1)},$$

and the latter are substituted into linearised load flow equations. The structure of these equations will not change, only diagonal blocks of Jacobian matrix and mismatches for generating nodes will be affected.

The algorithm has been programmed for digital computer using as a basis Newton's load flow program. The program is structured in such a way that additional static and dynamic elements of EPS can be easily introduced. The maximum size of the problem is about the same as in load flow calculations.

Test calculations for specific EPS has shown computational stability of the algorithm. For instance in one of the cases maximum error in δ (as compared to the results obtained for $\Delta t = 0.01s$) for integration step $\Delta t = 0.05s$ was equal to 2.6° , and for $\Delta t = 0.1s$ - 6° .

5. Conclusions.

1. The use of implicit numerical integration method allowed to compile EPS transient processes calculations program based on Newton's load flow solution.

2. Convolution integral with piecewise linear approximation of input signal can be successfully used to reduce both order and "stiffness" of the set of differential equations describing dynamic elements of the system thus allowing for easy account of various controlling devices.

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7. Appendix. The Calculation of Response of Linear Dynamic Elements to an Arbitrary Input.

Continuous input signal sampled at arbitrary time instants t_i ($\Delta t_i = t_i - t_{i-1} > 0$) can be approximated as

$$x(t) = x_0 \cdot 1(t) + \sum_{i=1}^n \Delta x_i(t),$$

where: x_0 - initial value of x ; $\Delta x_i(t)$ - the variation of x during i -th time interval specified by adopted form of approximation; $1(t)$ - unit step function.

In the case of stepwise approximation we have

$$\Delta x_i(t) = (x_i - x_{i-1}) \cdot 1(t - t_i), \quad i = 1, 2, \dots$$

and the output can be calculated as

$$y(t_n) = x_0 h(t_n) + \sum_{i=1}^n (x_i - x_{i-1}) h(t_n - t_i),$$

$i = 1, 2, \dots$

where: $h(t)$ - transient function of the element, i.e. response of the element to $1(t)$ input signal.

In the case of piecewise linear approximation we have

$$\Delta x_i(t) = (x_i - x_0) t \times 1(t - t_0) / \Delta t_i,$$

$$\Delta x_i(t) = \left[x_i - x_{i-1} \left(\frac{\Delta t_i}{\Delta t_{i-1}} + 1 \right) + x_{i-2} \frac{\Delta t_i}{\Delta t_{i-1}} \right] \times$$

$$\times (t - t_{i-1}) \times 1(t - t_i) / \Delta t_i, \quad i = 2, 3, \dots$$

and correspondingly

$$y(t_n) = x_0 h(t_n) + (x_1 - x_0) h'(t_n) / \Delta t_1 +$$

$$+ \sum_{i=2}^n \left[x_i - x_{i-1} (\Delta t_i / \Delta t_{i-1} + 1) + x_{i-2} \Delta t_i / \Delta t_{i-1} \right] \times$$

$$\times h'(t_n - t_{n-1}) / \Delta t_i,$$

$$\text{where } h'(t) = \int_0^t h(t) dt - \text{response of the}$$

element to $t \times 1(t)$ input signal.

For typical linear dynamic elements having transfer functions of first and second order the above responses can be easily calculated by recurrent expressions which acquire the simplest form in case of constant sampling interval, i.e. $\Delta t_i = \Delta t = \text{idem}$.

For instance, for aperiodic element with transfer function $w(p) = k / (1 + pT)$ and transient functions $h(t) = k(1 - \exp(-t/T))$, $h'(t) = k(t - T + T \cdot \exp(-t/T))$ we have

$$y(0) = x_0 h(0) = 0,$$

$$y(n\Delta t) = y[(n-1)\Delta t] f_1 + f_{2n}, \quad n = 1, 2, \dots$$

where: $f_1 = \exp(-\Delta t/T)$,

$$f_{2n} = x_{n-1} k f_1$$

- for stepwise approximation,

$$f_{2n} = k[x_n - x_{n-1} f_1] - kT(x_n - x_{n-1})(1 - f_1) / \Delta t$$

- for piecewise linear approximation.

GENERATOR MODELLING AND LOAD MODELLING FOR POWER SYSTEM TRANSIENT STABILITY STUDIES

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ABSTRACT

This paper reviews the problem of generator modelling and load modelling for power system transient stability studies. For the generator models, the paper reviews the different models identified by the "IEEE working group on the determination and application of synchronous machine models for stability studies". For the load models, the paper focuses mainly on various models used for the static and dynamic load. Simulation studies are performed using some of these models in order to compare the transient stability results.

The paper gives guidelines on which model to choose in order to obtain the most conservative transient stability results.

1. INTRODUCTION

Power system stability is of interest mainly to electric utilities and industrial plants with co-generation. Various components affect the stability limit of the power system namely:

- the inertia of the generators and the rotating loads
- the impedance of the transmission and distribution lines
- the internal reactances of the machines
- the additional elements generally used for system control and
- the load and its characteristics

Whenever any of these components is planned for installation, their effects on the overall system stability must be determined.

From the electric utilities point of view, it is extremely difficult to accurately determine the system stability limit because some of the above-mentioned components, especially the

load, are not easy to establish. Furthermore, it is not always possible to perform on-line tests to determine the system stability due to economic considerations and unprecedented damage which might occur to equipment. For these reasons, computer simulations are employed and accurate models for power system components are required.

There is, at present, various models developed for different types of generators and loads. These models range from the simplest to the most complex. The electric utilities are usually at ease to choose a generator model to use for stability studies but the same cannot be said for load models. This is primarily because utilities do not always know the type of load used by their customers, but if they do, they cannot obtain the load parameters easily. Thus utilities tend to use less complex load models. Hence it is important to ensure that the stability limit calculated, should be on the conservative side in order to allow for any errors in modelling the load.

This paper compares the transient stability results obtained by using simple voltage dependent load models such as constant admittance, constant current and constant power. The local and remote loads are also modelled by these models.

2. POWER SYSTEM NETWORK MODEL

The network used to compare the transient stability results using different generator and load models is given in [1]. This network is a benchmark used for comparative studies in the power system community. The network parameters and the simulation conditions are also given in [1].

3. GENERATOR MODELLING

The subject of generator modelling has received much attention in the past, and is well documented [2] to [5]. The most practical generator models, as identified by the "IEEE working group on the determination and application of synchronous machine models" [2] are listed in table 1.

Model 0.0 is the simplest model and represents only the constant flux linkage of the machine.

Model 1.0 represents the effect of the field circuit as well. The other models (model 1.1 to model 3.3) represent a varying amount of damping in the direct and quadrature axis due to damper windings and slots in the rotor.

Table 1: Different generator models

	no damper winding	one damper winding	two damper winding	three damper winding
constant flux linkage	MODEL 0.0			
Field circuit only	MODEL 1.0	MODEL 1.1		
Field circuit + one damper circuit		MODEL 2.1	MODEL 2.2	
Field circuit + two damper circuits			MODEL 3.2	MODEL 3.3

Figure 1 shows the transient stability results obtained by using some of these models and the results are summarised in table 2.

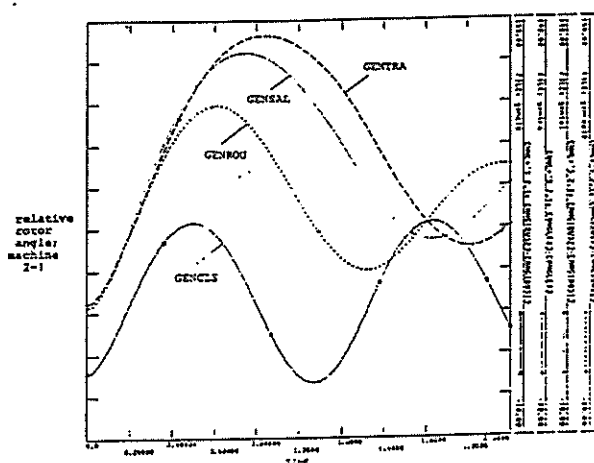


Figure 1 : Comparison of transient stability results using different generator models

Table 2 : Summary of results from figure 1.

Type of model	Transient stability results
Model 0.0	more optimistic
Model 1.0	more conservative
Model 2.1	conservative
Model 2.2	optimistic

Model 0.0 gives the most optimistic transient stability results. This model should be used only for the first swing transient stability study since it does not represent any damping in the generator.

Model 1.0 gives the most conservative transient stability results. This can be used for multiswing transient stability studies especially when the generator parameters are not known in detail.

Model 2.1 gives conservative results. The model can be used both in transient stability and dynamic stability studies since it represents the limited damping effect of the generator. This model is adequate for modelling hydraulic generators.

Model 2.2 gives optimistic transient stability results. This model can also be used for both transient and dynamic stability studies since it represents more damping of the generator circuits. This model is adequate for turbogenerators without damper windings.

4. LOAD MODELLING

Power system loads are classified into two categories, the static load and the dynamic load. Static loads are passive or non-rotating loads. They can be reasonably expressed by polynomials or some form of algebraic functions like exponentials. They can either be voltage dependent, frequency dependent or both. The various static load models are discussed briefly below.

i). *Constant admittance model*

The complex power varies as the square of the voltage to keep the admittance constant. The equations expressing this representation can be written as

$$Y = (P - jQ) / |V|^2 = \text{const} \dots\dots\dots(1)$$

where Y represents the equivalent shunt admittance of the load.

ii). *Constant current model*

The complex power varies proportionally with the voltage to keep the load current constant. The equation expressing this representation is given by

$$I = (P - jQ) / V^* = \text{const} \dots\dots\dots(2)$$

where I represents the load current.

iii). *Constant power model*

The complex power is assumed to remain constant during and after the disturbance. The equation expressing this representation is given by

$$S = V \times I^* = P + jQ = \text{const} \dots\dots\dots(3)$$

Dynamic loads, on the other hand, are rotating loads whose output varies considerably and rapidly to changes in voltage and frequency. The most common example of a dynamic load is the induction motor and specific models are described in [6] to [8].

Because of the problems of obtaining accurate data (parameters) for dynamic loads, static load models are often used in place of dynamic loads. This approach may lead to incorrect transient stability results. It is important to ensure that the results obtained by this approach are conservative.

Figure 2 shows the transient stability results obtained by using different static load models. The figure shows the relative rotor angle of two machines as a function of time. Modelling the load as constant power gives the most conservative transient stability results, while modelling the loads as constant admittance gives the most optimistic results.

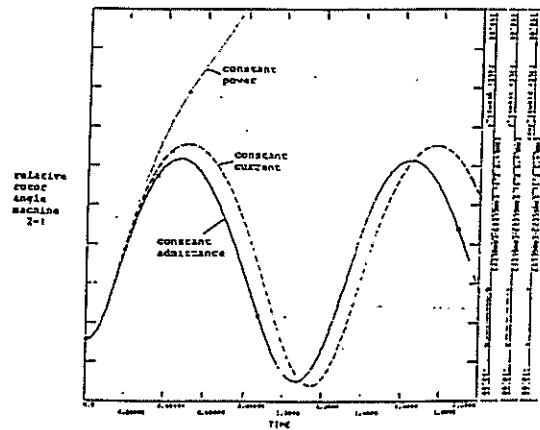


Figure 2 : A comparison of different static load models

From figure 2, one is inclined to draw a conclusion that the constant power model always gives the most conservative result. However, looking at figure 3, the results are the most conservative when the local loads are modelled as admittance.

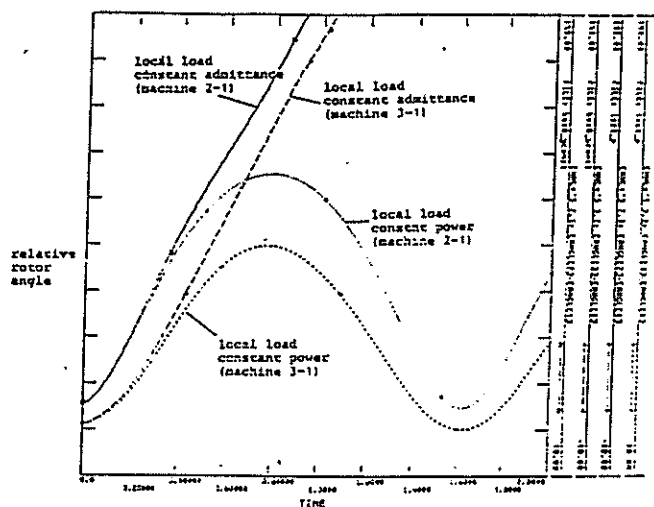


Figure 3: A comparison of different static load models for a local load

Figure 4 shows the transient stability results of the remote loads modelled as constant power and constant admittance load. The results in this case are conservative when the remote load is modelled as constant power.

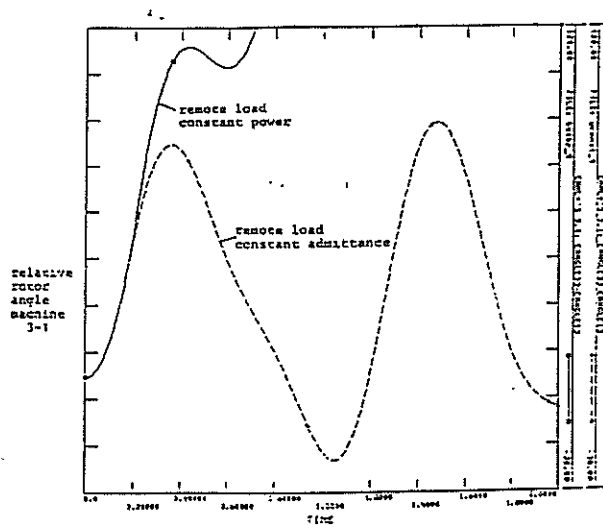


Figure 4: A comparison of different static load models for a remote load

The different results in figures 3 and 4 is due to the fact that during a major fault, the synchronous machines either accelerate or decelerate depending on their location with respect to the load and the fault. A summary of the transient stability results obtained by using different static load models is listed in Table 3.

Table 3: Summary of load modelling simulations

Load model	All loads	Local load	Remote load
constant power	conservative	optimistic	conservative
Constant current	optimistic	conservative	optimistic
Constant admittance	more optimistic	more conservative	more optimistic

5. CONCLUSIONS

1. When using static load models such as constant power, constant current and constant admittance, the relationship between the load model, the location of the fault and generation should be considered.

2. Modelling a local load as constant admittance gives the most conservative transient stability results than when the load is modelled as constant power. Thus a local load should be modelled as constant admittance.

3. Modelling a remote load as constant admittance gives the most optimistic transient stability results than when the load is modelled as constant power. Thus a remote load should be modelled as constant power.

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DYNAMIC STABILITY INVESTIGATIONS -- ON THE SELECTION OF NEGATIVE DAMPING STABILIZER AND GOVERNOR PARAMETERS

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Self-excited oscillations are likely to be exhibited, under certain conditions, by dual-excited synchronous machines connected to series capacitor compensated long transmission systems. This paper deals with study of the effect of negative damping stabilizer (NDS) and the combined action of NDS and fast-acting governors on these oscillations in a hydrogenerator system, taking into account the torsional dynamics of the turbine-generator system. A procedure for the proper selection of the parameters of the excitation controller and of the governor is presented.

1. INTRODUCTION

Although much interest has been shown in the past for dynamics and transient stability studies of dual-excited synchronous machines, very little work has been reported regarding self-excited oscillations of these machines in a series capacitor compensated long distance transmission systems [1,2].

El-Serafi et al. [1] have examined the problem of self-excited oscillations for unregulated dual-excited synchronous machine systems. Some work has been reported by Hurry et al. [2] on the effect of excitation control of the field windings of a dual-excited synchronous generator connected to an infinite busbar through a series capacitor compensated transmission line. A current regulator in the reactive field winding was found to be effective in suppressing the low-frequency oscillations only, whilst the simultaneous control of the reactive and torque field windings was found to be inferior to that of an equivalent conventional synchronous machine as far as suppression of self-excited oscillations are concerned. Saito et al. [3] showed that an NDS can suppress high frequency self-excited oscillations in a conventional synchronous machine connected to capacitor compensated transmission line. However, the system considered by the authors does not include the turbine generator dynamics and the analysis was restricted to the high frequency oscillations only.

It was reported [4] that mechanical governors are unable to improve the subsynchronous resonance stability limits whereas properly designed electrohydraulic governors are able to do so. The effect of an electrohydraulic governor on self-excited oscillations has been studied in the past [4,5].

However, these studies pertain to systems with conventional synchronous machines. Besides, in these studies the governor was assumed to be directly attached to the generator and not to the turbine as it is the case in practice.

Hydroturbines introduce negative damping in power system, sometimes giving rise to instability. It has been demonstrated [6] that compensation schemes for hydroturbine units improve the transient stability of power systems. However, no work has been reported to the effect of such schemes on the self-excited oscillations in systems with dual-excited synchronous machines. In view of this, the author is motivated to investigate the effect of compensation schemes on self-excited oscillations.

This paper presents an attempt to study the self-excited oscillations phenomenon of dual-excited synchronous machines in hydrosystems taking into account the torsional dynamics of the turbine-generator system. The studies pertain to the design aspects of the NDS and governor that can suppress the self-excited oscillations for different hydrosystem schemes.

Stability studies are presented for the following cases :

- (a) the unregulated system,
- (b) the generator with its reactive field winding regulated by an NDS whilst its torque field winding is left unregulated,
- (c) the hydrogenerator with simultaneous control of both its reactive and torque windings, and
- (d) the hydrosystem with parallel and feedback compensation schemes.

2. SYSTEM INVESTIGATED

The system chosen for the study reported in this paper is shown in Figure 1. It consists of a dual-excited synchronous machine connected to an infinite busbar through to series-compensated transmission line.

All investigations are carried out with the system parameters given in the companion paper [7] and the following:

$X_d = 1.2$ p.u.

Electrohydraulic governor :

$T_1, T_3 = 0.02$ s ; $T_2 = 0$ s ; $T_4 = 0.5$ Tw = 0.5 s ; $K_T = -2$

Mechanical speed governor :

$T_1 = 48.7$ s, $T_2 = 5$ s, $T_3 = 0.513$ s, $T_4 = 0.5$ s ; $K_T = -2$

3. ANALYSIS

The small displacement model for self-excited oscillations for the hydrosystem when both the reactive and the torque field windings of the synchronous machine are regulated can be written in matrix form as

$$[A] \cdot [\Delta x] = 0 \quad (1)$$

where $[A]$ is a 5×5 matrix whose elements are defined in the Appendix, and

$$\Delta x = [\Delta v_r \ \Delta v_t \ \Delta i_d \ \Delta i_q \ \Delta \delta]^T \quad (2)$$

In this model, the torque field winding is assumed to be actuated by an angle regulator whereas the reactive field winding is regulated by an NDS.

3.1 Case I : The unregulated system.

The small displacement model for the unregulated system is obtained from equation (1) by setting

$$\Delta v_r = \Delta v_t = 0 \quad (3)$$

The corresponding characteristic equation is obtained by equating the principal determinant of the resulting matrix to zero.

Figure 2 shows the dynamic stability limits for the system. From similar work reported for the case of a conventional synchronous machine [7], it is observed that the dual-excited machine is inherently more stable than its equivalent conventional one. As the level of excitation is increased, instability due high frequency self-excited oscillations occurs at higher values of the armature resistance

The stability limit is affected by the value at which the rotor angle is fixed. From eigenvalue analysis, it is found that for a certain fixed value of the armature circuit resistance, the frequency of the low frequency oscillations increases with increase in the rotor angle and level of excitation. Higher values of the two field winding resistances tend to decrease the frequency of these oscillations.

3.2 Case II : NDS on the reactive field winding

For the present case of interest, the reactive field winding is regulated by an NDS whilst the torque field winding is unregulated. The corresponding displacement model is obtained by eliminating the elements of the second row and second column from the matrix $[A]$. Parameter plane technique is used for the investigations reported herein.

From Figure 2, it is evident that operating points A ($R, X_C = 0.6, 1.0$) and B ($R, X_C = 0.05, 0.9$) are unstable for the unregulated system. Results of analysis (Figures 3,4) reveal that a proportional NDS is effective in suppressing both low frequency and high frequency self excited oscillations. Stability limit curves are also shown for different degrees of damping (λ).

The incorporation of additional signals (Figure 5) in the form of the first and second derivatives of the reactive power signal is seen to extend the stable zones further, by providing stable operation for

practical values for both the proportional gain (0.5) and the exciter time constant T_{er} (0.02 s).

Better degrees of damping (λ) can still be achieved through the use of higher values of only the first NDS derivative gain i.e. $K_{2n} = K_{0n} = 0$.

Both mechanical speed and electrohydraulic governors (EH) are found to decrease both the size of stable regions and the degree of stability obtained by proportional NDS regulators (Figures 6,7).

In Figure 5, the point corresponding to $K_{1n} = 200$ and $K_{2n} = 8000$ is unstable. Analysis (Figure 8) reveals that the incorporation of the first and second derivatives of the speed signal in the governor-turbine loop can make this point stable. An increase in the value of the proportional governor gain has an adverse effect on the size of the stable region.

3.3 Case III : Simultaneous control of field windings.

The performance of the machine when both field windings are regulated is found to be inferior to that of the machine regulated by the NDS operating alone.

3.4 Case IV : Hydroturbine Compensating Schemes

The compensating schemes of Ref [7] are used for investigations reported in this section.. Results of investigations (Figure 9) reveal that parallel compensation of the turbine systems can improve the stability in the presence of both mechanical and electrohydraulic governors. Further analysis shows that an increase in the compensating network's gain reduces the size of the stable region in the K_{2G}/K_{1G} plane in the presence of derivative NDS gains.

Figure 10 indicates that feedback compensated hydroturbine systems can equally improve these stability limits. Likewise, an increase in the gain of the compensating network decreases the stable region in the $K_{2G}-K_{1G}$ plane.

4. CONCLUSIONS

The parameter plane technique has been effectively used to investigate the combined effect of negative damping stabilisers and governor action on subsynchronous resonance. This allows for a convenient procedure for the proper selection of the controller and governor parameters to ensure suppression of self excited oscillations.

Both parallel and feedback turbine compensation can also suppress low frequency self excited oscillations.

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6. NOMENCLATURE

δ_L, δ_L rotor and load angles
 v_r, v_t reactive and torque field winding voltages
 i_r, i_t reactive and torque field winding currents
 v_d, v_q d- and q- axis components of voltage
 Ψ_r, Ψ_t flux linkages in reactive and torque windings
 X_d', X_q' transient reactance in d- and q-axis circuits
 X_{ad}, X_{aq} Mutual reactances in d- and q-axis circuits
 X_r, X_t total reactance in reactive and torque windings
 $K_r(p), K_t(p)$ transfer functions of regulators in reactive and torque windings
 R_r, R_t rotor resistances in reactive and torque windings
 $X_q(p)$ q-axis operational impedance
 $G_d(p), G_q(p)$ d- and q-axis operational functions
 X_{aq} q-axis magnetizing reactance
 λ degree of damping $p = -\lambda + jw$
 w angular frequency
 T_{er}, T_{et} exciter time constants for reactive and torque field windings
 T_a, T_b rectifier time constants for reactive and torque field windings
 T_{ro}, T_{to} open-circuit time constants for reactive and torque field windings
 H inertia constant (no torsional dynamics)
 K_{on}, K_{1n}, K_{2n} proportional, first and second derivative gain
 $K_{o\delta}, K_{1\delta}, K_{2\delta}$ of reactive power feedback signal

proportional, first and second derivative gain
 of angle regulator
 H_g, H_{T1} Turbine-generator inertia constants
 K_{T1g} Turbine-generator shaft stiffness
 T_w water inertia constant
 T_1, T_2, T_3, T_4 governor constants
 $g_D(p)$ governor transfer function
 $X'_{d\Sigma} = X'_d + X_e$
 $X'_{q\Sigma} = X'_q + X_e$
 $X_d(p) = (pT_{ro}X'_d + X_d) / (pT_{ro} + 1)$
 $X_q(p) = (pT_{to}X'_q + X_q) / (pT_{to} + 1)$
 $X_{d\Sigma}(p) = (pT_{ro}X'_{d\Sigma} + X_{d\Sigma}) / (pT_{ro} + 1)$
 $X_{q\Sigma}(p) = (pT_{to}X'_{q\Sigma} + X_{q\Sigma}) / (pT_{to} + 1)$
 $G_d(p) = X_{ad}/R_r$
 $G_q(p) = X_{aq}/R_t$
 G_c turbine compensation constant

Other symbols are given in companion paper [7].

7. ADDRESS OF AUTHOR

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8. APPENDIX

Elements A_{ij} 's of the model in equation (1):

$$\begin{aligned}
 A_{11} &= 1 & A_{12} &= 0 \\
 A_{13} &= -K_r(p) M(p) \\
 A_{14} &= -K_r(p) N(p) & A_{15} &= -K_r(p) L(p) \\
 A_{21} &= 0, A_{22} = 1, A_{23} = 0, A_{24} = 0, A_{25} = -K_t(p) \\
 A_{31} &= (1-p^2)G_d(p) & A_{32} &= 2pG_q(p) \\
 A_{33} &= p^2X_d(p) - X_d(p) + pR_\Sigma + X_C \\
 A_{34} &= -(2pX_d(p) + R_\Sigma) \\
 A_{35} &= p^2\Psi_{qo} + 2p\Psi_{do} + v_{do} - p i_{qo}R_\Sigma \\
 A_{41} &= -2pG_d(p) & A_{42} &= (1-p^2)G_q(p) \\
 A_{43} &= (2pX_d(p) + R_\Sigma) \\
 A_{44} &= p^2X_q(p) - X_q(p) + pR_\Sigma + X_C \\
 A_{45} &= -p^2\Psi_{do} + 2p\Psi_{qo} + v_{qo} + p i_{do}R_\Sigma \\
 A_{51} &= i_{qo}G_d(p) & A_{52} &= -i_{do}G_q(p) \\
 A_{53} &= -(\Psi_{qo} + i_{qo}X_d(p)) \\
 A_{54} &= \Psi_{do} + i_{do}X_q(p) \\
 A_{55} &= H p^2 + g_D(p) \text{ (modified to include torsional dynamics)} \\
 H &= H_G p^2 + (D_{GG} + D_{TIG})p + K_{TIG} - (D_{TIG}p + K_{TIG})^2 / [H_{T1}T_2 p^2 + (D_{T1}T_1 + D_{TIG})p + K_{T1G}] \\
 g_D(p) &= [K_{0g} + pK_{1g} + p^2K_{2g}] [pT_4K_r + 1] (pT_2 + 1) / [pT_1 + 1] (pT_3 + 1) (pT_4 + 1)
 \end{aligned}$$

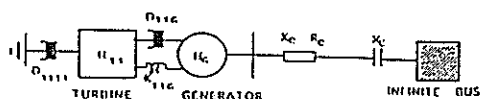


Figure 1: System investigated.

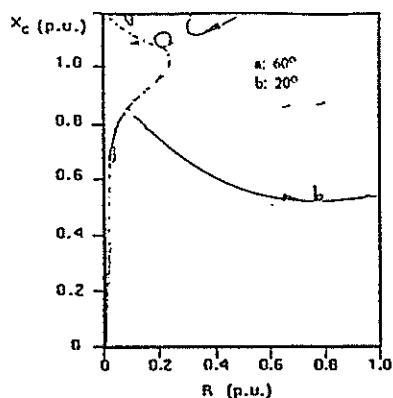
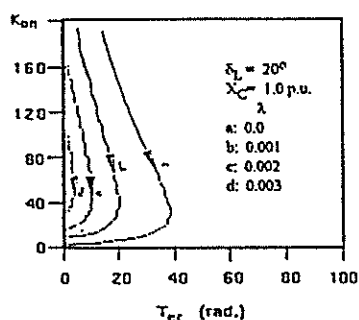
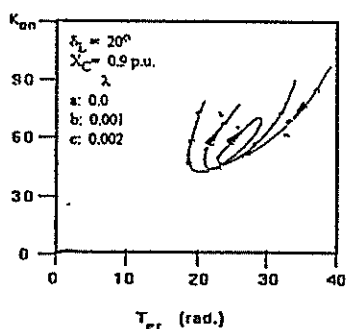
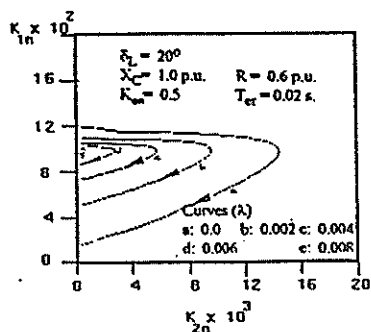
Figure 2: Stability limit curves ($\delta_T = 0^\circ$).Figure 3: Stability zones (no governor) $R=0.6$ p.u.Figure 4: Stability zones (no governor) $R=0.05$ p.u..

Figure 5: Effect of derivative NDS gains.

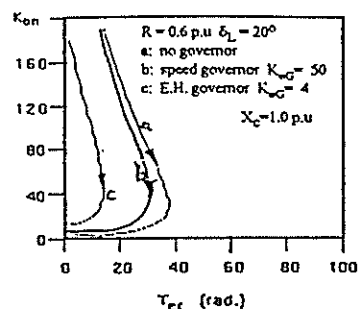


Figure 6: Effect of proportional governor gain.

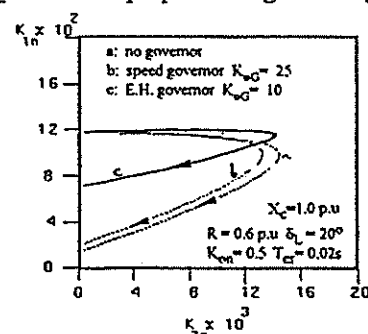


Figure 7: Effect of governor/NDS gains.

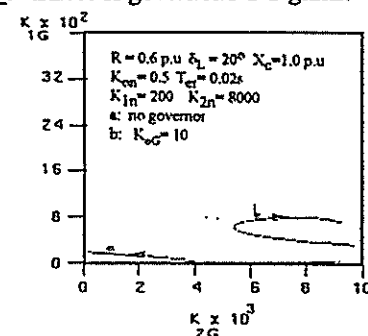


Figure 8: Effect of governor derivative gains.

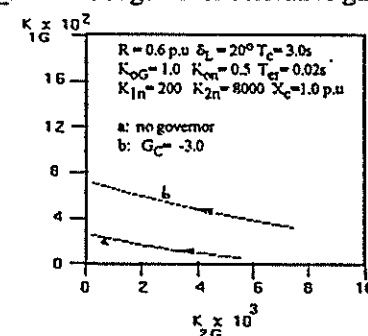


Figure 9: Parallel compensation effect.

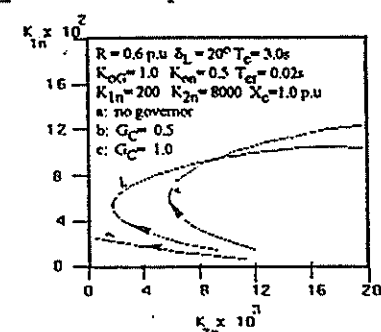


Figure 10: Feedback compensation effect.

PV MAXIMUM POWER POINT TRACKER: DESIGN OF EFFICIENT CONVERTER AND SIMPLE CONTROLLER

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ABSTRACT - A 20% cost reduction of photovoltaic generation can be brought about, by the investment in a novel, highly efficient maximum power point tracker (MPPT). Such a converter is currently under development, implementing a feedforward output current controller as tracking algorithm. The dc-to-dc converter, typical of an MPPT, exploits the efficient high-frequency operation, associated with zero-transition switching. Thus the output filter only consists of the combination of a storage battery and its cabling. The resulting converter is suitable for integration into standard PV panels during manufacturing. The efficiency of the MPPT is 96% for a range of output powers at a low battery voltage of 12V and a component cost of US\$33,82.

1. INTRODUCTION

The costliest per-Watt-expenditure in a photovoltaic system arises in the generation. The current-voltage and power-voltage characteristics of a solar panel change with the meteorological conditions, which the panel is exposed to. The typical battery load, however, requires a constant voltage which improbably corresponds to that of the array at its maximum output. It is the function of a maximum power point tracker (MPPT), to adjust its input voltage, which is also the array voltage, so that it corresponds to the maximum power output of the panel. At its output the MPPT always provides the constant voltage required by the load. Thus an MPPT requires a dc-to-dc converter, controlled by a maximum power tracking algorithm. Experiments performed imply, that an MPPT with an efficiency of 85% can increase the amount of energy stored in batteries on average by 17% per solar day [3].

The proposed economisation is implemented by the construction of an MPPT of extreme efficiency and minimal price. Such a construction is made possible by using a very economic dc-to-dc converter.

Since filter inductors and capacitors are especially costly with increasing size, the approach was reducing their size by choosing a sufficiently high converter switching frequency. Simultaneously the high efficiency was obtained by selecting an extremely simple, soft-switched topology out of several [6], [5], [2]. The simplicity is advantageous, because with each power switch driver circuitry and losses are associated. The traditional freewheeling diode was in this converter substituted by a reverse conducting MOSFET, (T_D in Fig. 1) since its conduction losses were appreciably lower than those of typical diodes [7].

2. DETAILS OF THE SOFT-SWITCHED DC-TO-DC CONVERTER

The requirements for the converter, formulated in paragraph 1, are met in the topology depicted in Fig. 1. Normally this topology is shown with an extra switch in series with the snubber capacitor C_s [6]. The switch can be left out, since the turn-off snubber is recovered by the non-linear inductor L_N - the exact detail is given in the cycle description below. Its simultaneous simplicity and efficiency make it ideal for the MPPT-application. The synchronous rectifier, (T_D) cable (L) and battery as external output filter are shown. The positive output current feedback, for the purpose of maximum power point tracking, is also schematically presented in Fig. 1.

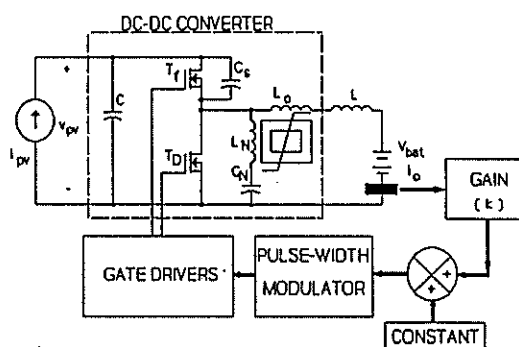


Fig. 1: Schematic of chosen topology with feed-forward MPPT control

(a) Description of the converter cycle

The converter cycle is described in lettered intervals corresponding to the circuit diagrams in Fig. 2 and the waveforms in Fig. 3.

a. Initially the converter is freewheeling and the channel of MOSFET T_D is on. It then simply acts as a silicon resistor. L_N and C_N in Fig. 1 act as lowpass filter, with the consequence, that C_N has long since charged to approximately the output voltage, V_{bat} . V_{bat} is thus impressed across L_N in this time period, resulting in a rise of current through this coil. The output current I_o through L_o provides a magnetomotive bias in the core of the non-linear inductor, so that saturation occurs at the end of this interval. Consequently the current in L_N rises to load-current level.

b. This is the latest instant, on which the channel of the synchronous rectifier can be switched off.

c. The current in the non-linear inductor, i_{L_N} , rises above the output current, recovering the parasitic diode of the synchronous rectifier, T_D , as shown in Fig. 2. Simultaneously the drain-source capacitance of T_D is

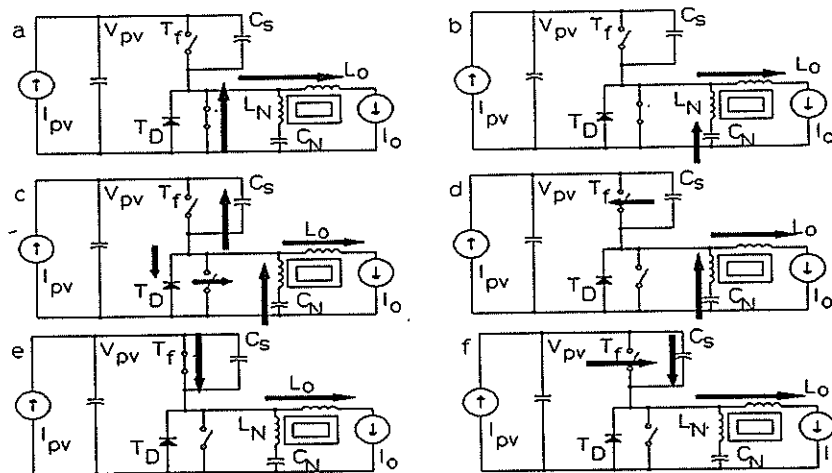


Fig. 2: Pictorial depiction of converter cycle

charged and the snubber capacitance C_s (in the same figure) is discharged.

d. After a successful commutation described in instant c, the snubber capacitance is completely discharged, while the load current is carried by the non-linear inductor L_N in Fig. 2. The MOSFET T_f can now be switched on losslessly.

e. This is the conduction cycle of the MOSFET, T_f . The current in the output inductance rises slowly.

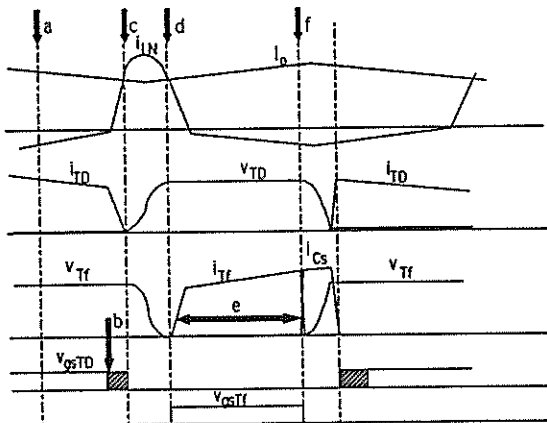


Fig. 3: Waveforms corresponding to previous pictorial representation.

f. At turn-off of the main MOSFET (T_f in Fig. 2), the snubber capacitor is charged by the load current commuting through it - view i_{Cs} in Fig. 3. The turn-off of the MOSFET is thus nearly lossless. The parasitic diode of the synchronous rectifier switches on as the snubber capacitance is fully charged. The channel of the synchronous rectifier can be switched on any instant after period f has elapsed.

If Fig. 3 is studied, it becomes evident, that the converter operation closely resembles perfect pulse-width

modulation, since the time consumed by the action of the snubber and the non-linear inductor are negligible compared to the switching period. The saturation period of the inductor has been drawn out of proportion in Fig. 3 to facilitate the explanation.

Such a converter was constructed. In Fig. 4, the voltage across and the current through the synchronous rectifier,

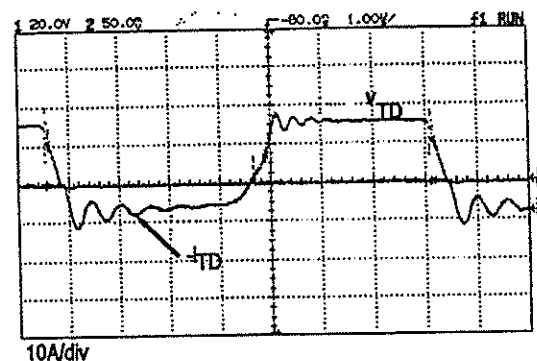


Fig. 4: Voltage across and current through T_D .

T_D , are shown. Soft switching can clearly be observed in the figure. In Fig. 5 the turn-on of T_f is shown. After the discharge of the snubber capacitance C_s by the non-linear inductor current, one can observe the drain source-voltage of T_f , v_{TF} to become slightly negative. This occurs, because the excess inductor current then flows through the body-drain diode of T_f . Only after the gate voltage of a MOSFET has gone high, its drain current starts flowing. Since the gate voltage only goes high, after the voltage across the MOSFET T_f is low, the depicted turn-on is soft.

3. THE FEEDFORWARD MPPT CONTROLLER

In this paragraph, the controller necessary to make an MPPT out of the presented dc-to-dc converter, is

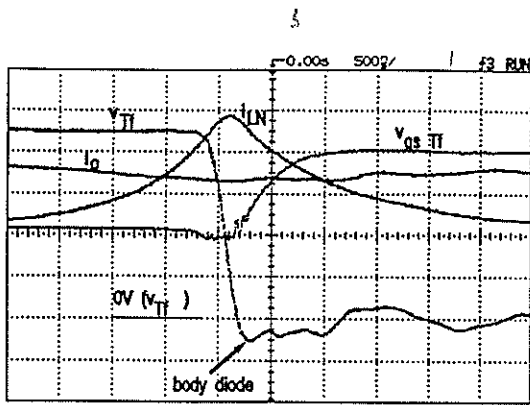


Fig. 5: Prototype-measured waveforms

designed. The theory of this controller is well documented [4], [8]. Therefore this paper concentrates more on the design and the results.

Since the output battery has a voltage that varies with only a small percentage due to its changing state of charge, one can describe the output power as follows:

$$\begin{aligned} P_o(t) &= V_{bat} i_o(t) \\ P_o(t) &\propto i_o(t) \end{aligned} \quad (1)$$

where:

P_o i_o = the output power and current

V_{bat} = the constant battery voltage

Thus, when assuming the battery voltage to be constant, the output current is a proportional reflection of output array and converter power.

In Fig. 1 and Fig. 7, the feedforward controller is shown: it consists of a simple gain between the output current and the control voltage determining the duty cycle. The typical control action of the feedforward controller functions like this: Initially, the input capacitor C is charged to open-circuit array voltage (position 1 in Fig. 6). As the output current increases, so does the duty cycle. Due to the constant-gain feedback more current is drawn out of the capacitor C in Fig. 7, than is supplied by the array (i_{pv}). The input voltage of the MPPT decreases, while output current and duty cycle increase.

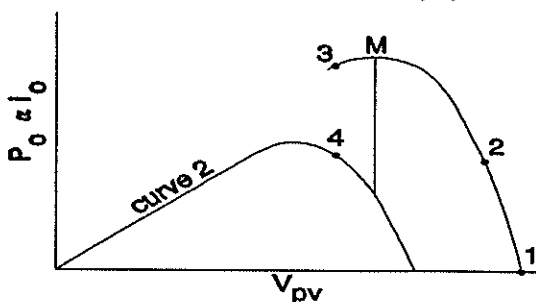


Fig. 6: Curves, illustrating principle of feedforward controller

If an overshoot occurs, the controller reaches position 3 in Fig. 6. Since the current drawn by the converter is now lower than the photovoltaic current, the capacitor C

is charging, with the result that finally, the converter reaches point M in Fig. 6. In similar fashion, the maximum of curve 2 in Fig. 6 is reached, which becomes valid after a sudden change in meteorological conditions. One can also follow this argument by studying the block diagram of the control problem in Fig. 7. Since the converter switching frequency is much higher than the filter dynamics, while the ripple currents and voltages are very small, one can represent the converter voltages and currents in terms of averages. A block diagram was thus synthesised, with whose aid one can also study the dynamics of the feedforward controller.

The design is straightforward. Since a limited range of

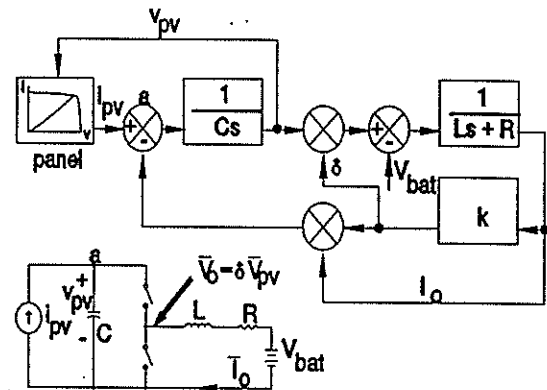


Fig. 7: Block diagram identifying control problem

output currents is expected for one particular converter, one can, after having constructed a low-impedance shunt, identify the expected shunt voltage range. Practical limits also exist for the duty cycle range (in this case 10 to 85%), which can also be expressed in control voltages. Thus one simply uses operational amplifiers, to convert the given shunt voltage range into the given control voltage range.

Due to the constant gain feedback of output current, one can write:

$$\delta(t) = k i_o(t) \quad (2)$$

where:

δ = the duty cycle

i_o = the output current

k = the feedback gain

Looking at Fig. 7, one notes, that the steady state output current can be written as follows:

$$\frac{\delta \bar{V}_{pv} - V_{bat}}{R} = \bar{I}_o \quad (3)$$

where:

$\bar{V}_{pv}$ $\bar{I}_o$ = the steady state values of array voltage and output current

Using equation (2) in the steady state:

A steady-state settling locus results for the feedforward controller. It takes the form of the hyperbola shown in

$$\frac{k\bar{I}_o \bar{V}_{pv} - V_{bat}}{R} = \bar{I}_o$$

$$\text{rearranged: } \bar{I}_o = \frac{V_{bat}}{k\bar{V}_{pv} - R} \quad (4)$$

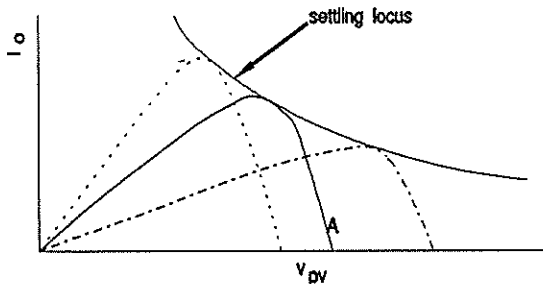


Fig. 8: Settling locus, on which the feedforward controller performs optimally

Fig. 8. In nature, the array power maxima shift due to insolation and temperature in a hyperbola too [1]. For optimum controller operation, the gain of the current feedback, k , in Fig. 7, must be set in such a way, that the settling locus closely corresponds to the hyperbola, according to which the array power curves shift in nature.

4. MPPT PERFORMANCE

The dc-to-dc system described in paragraph 2 had the efficiency shown in Fig. 9. Undue resistance between the converter output and the battery limited it. This resistance was measured, and the possible efficiency of the system was determined, and also plotted in Fig. 9. Conclusively, it can be said, that the converter efficiency was very high; the exact percentage of the system depends on the improvement of the installation and a remeasurement.

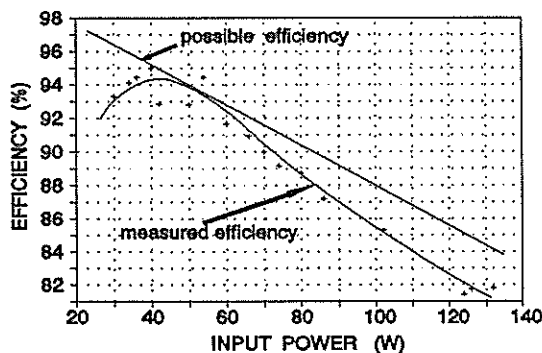


Fig. 9: Open-loop efficiency obtained with the MPPT

In Fig. 10, the performance of the MPPT controller is given. Black dots and asterisks indicate the maxima perceived by the controller. The curves with the black dots were taken with one gain k , while those with asterisks were taken with another. Curve A was taken at 10:20 while curve B was taken at 15:45.

The total cost of the MPPT system was US\$33,82. A cost decrease of 16,6% arises in the price per installed 165W of photovoltaic generation.

5. CONCLUSIONS

The principle was demonstrated, that the introduction of an inexpensive, yet efficient MPPT to a typical PV system is a benefit. Without doubt, the manufacturability of the proposed converter can be improved; this has, however, to take place in co-operation with the industrial engineers responsible for the manufacture.

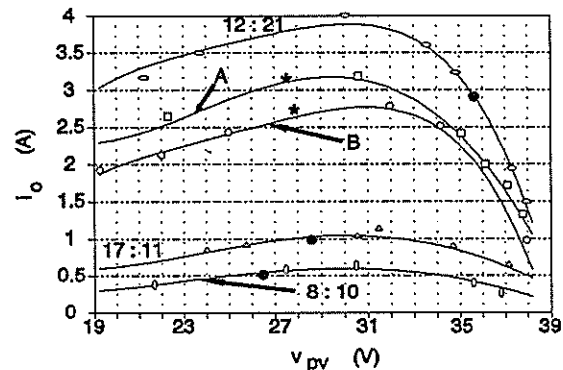


Fig. 10: Accuracy achieved with feedforward MPPT controller

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AC TO DC CONVERTER WITH UNITY POWER FACTOR AND MINIMAL HARMONIC DISTORTION

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ABSTRACT : Conventionally used AC to DC converters such as the three phase diode and phase-controlled thyristor stacks are notorious for the high levels of harmonic rich line currents that they draw from the three phase mains supply. In addition, the thyristor stack presents a poor power factor to the AC supply for low values of DC output voltage. This paper proposes a solution to the harmonic and power factor problems by the use of a new force-commutated rectifier topology which when appropriately controlled is capable of achieving the following objectives:

- (a) unity input power factor,
- (b) near sinusoidal line currents and
- (c) bidirectional power flow

without the need for large AC filters and reactive power compensators.

1. INTRODUCTION

Diode and phase-controlled thyristor rectifiers are in wide use today in applications such as dc and ac variable speed drives, battery chargers, uninterruptible power supplies, arc welding equipment, and HVDC transmission systems. However, these devices possess some serious shortcomings.

In the case of the diode stack, bidirectional power flow is not possible, and the input line currents to the rectifier are rich in harmonics. Thyristor rectifiers are also characterised by harmonic rich line currents. In addition, the power factor worsens for decreasing output voltages.

The distorted line currents drawn by these devices has meant that their widespread use has increased the danger of severe harmonic distortion of the ac power supply network. Thus far, distortion of the ac supply has been avoided by the installation of ac harmonic filters. In addition, reactive power compensators are used to improve the power factor of thyristor rectifiers. In recent times researchers have been investigating improved rectifier structures which when appropriately controlled, would possess improved power factors and near sinusoidal line currents, and in so doing obviate the need for large filters and reactive power compensators. These new devices have been termed 'force-commutated reversible rectifiers'.

Like inverters, these force-commutated rectifiers can be classified into two categories according to the nature of the dc side [1]. Voltage-sourcing devices are characterised by a filter capacitor across the dc link and filter inductors in each ac supply line. Current-sourcing devices are characterised by filter inductors in series with the dc link to give the device a constant current characteristic, and filter capacitors on the AC side. This paper discusses a voltage-sourcing device which is suitable for use in VSI type drives. The operating principles are discussed after which simulation results are presented.

Figure 1 shows a typical voltage sourcing force-commutated reversible rectifier which is capable of achieving bidirectional power flow, near sinusoidal line currents and unity input power factor. As can be seen,

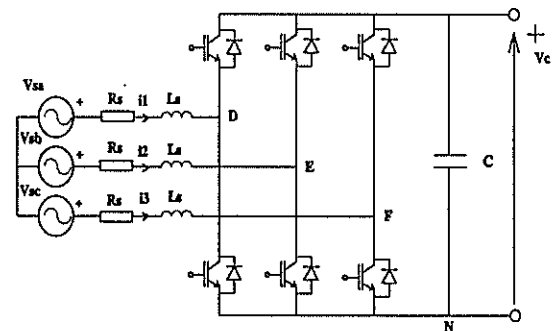


Figure 1 : Three Phase Force-commutated Rectifier

it consists of six force-commutated switches and six antiparallel connected diodes. The IGBTs shown on the diagram may be replaced by any other force-commutated switch such as the MOSFET, Power Transistor, MOS-controlled Thyristor or GTO, depending on the application. Each switch module is capable of carrying current in both directions and blocking voltage in one direction only making the converter two-quadrant. Regenerative operation is achieved by reversing the flow of the direct current back into the converter. Comparison of this converter with the voltage source inverter (VSI) reveals that the topologies are identical so that the converter is completely reversible and can be made to operate as an inverter very easily.

There are two fundamental approaches that are used to control the converter shown in figure 1. One is referred to as the voltage-forcing approach while the other is termed the current-forcing approach [1]. The voltage-forcing strategy involves generating a set of balanced three phase sinusoidal PWM voltages between the converter arm midpoints and the star point of the supply. By controlling the magnitude and phase of the fundamentals of these voltages, the phase and magnitudes of the line currents can be regulated, and in so doing the dc link voltage can be regulated. This paper details a current forcing control scheme termed Hysteresis Current Control (HCC) in which the actual line currents are forced to track sinusoidal reference waveforms. When compared with many of the voltage forcing schemes e.g the one discussed in [3], HCC has the following advantages:

- (a) excellent transient response,
- (b) extremely simple controller hardware,
- (c) adaptability to nonlinearities such as transistor switching delays.

The main disadvantages of this scheme are the variable switching frequency and the fact that the current harmonics are not readily definable.

2. CURRENT-FORCING PRINCIPLE

In order to understand the current-forcing principle, consider the single phase case shown in figure 2.

Assume that the capacitors have charged up initially through the inverse parallel diodes without any transistor switching having taken place. Each diode in the converter acts as a freewheeling diode for the opposite transistor in the same leg. For example, if transistor T1 is conducting and is then turned off by the application of a negative gate to emitter voltage, the current that it was conducting instantaneously freewheels through diode D2 due to the inductive nature of the circuit. In a similar manner, diode D1 acts as the freewheeling diode for transistor T2. Each

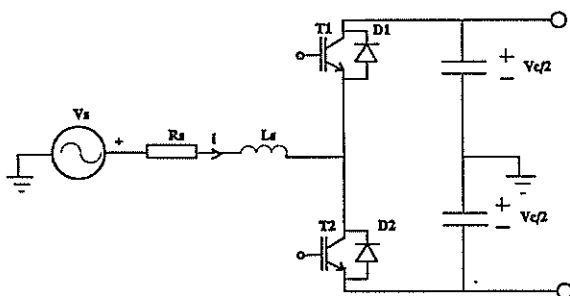


Figure 2: Single Phase Rectifier

transistor-diode pair therefore acts as a fully controlled bidirectional switch. It is therefore useful to replace each switch pair by an equivalent, single bidirectional switch.

Then, if the lower switch is conducting the following equation may be written:

$$L_s \frac{di}{dt} + R_s i = V_{sa} + \frac{V_c}{2} \quad (1)$$

It is required that the capacitor voltage $V_c/2$ always be greater than the source voltage. If this condition is satisfied, the RHS of (1) will be positive (neglecting the effect of R_s) so that if the lower switch is conducting (i.e either T2 or D2 is conducting), the current will ramp up (positive di/dt). Similarly, when the upper switch is conducting (i.e either T1 or D1) the following equation can be written:

$$L_s \frac{di}{dt} + R_s i = V_{sa} - \frac{V_c}{2} \quad (2)$$

Once again, if $V_c/2$ is greater than the source voltage, the RHS of (2) will be negative meaning that the current will ramp down when the upper switch is on (since di/dt is negative). It is therefore apparent that the appropriate switching of transistors T1 and T2 enables one to force the line current in a desired direction. In a practical implementation this capability is used to force the line current to track a sinusoidal reference waveform which is in phase with the source voltage in order to maintain unity power factor. In order to limit the switching frequency, a hysteresis band is defined about the sinusoidal reference waveform as depicted in figure 3. If the line current exceeds the upper hysteresis boundary, the lower transistor is switched on and the upper transistor switched off and if the line current goes below the lower hysteresis boundary, the lower transistor is switched on and the upper transistor is switched off.

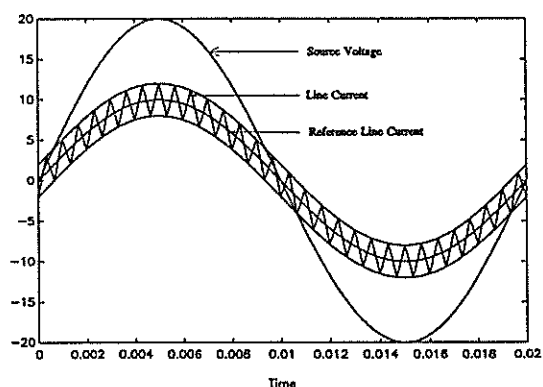


Figure 3: HCC Principle

Figure 4 shows how the transistor switching is controlled according to this principle.

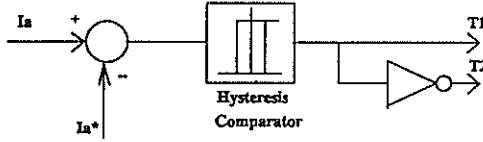


Figure 4: Control of Transistor switching

3. THREE PHASE OPERATION

In order to obtain a clearer understanding of how the current-forcing principle works in the three-phase case, it is useful to examine the mathematical model of the system. Each of the transistor/diode pairs is represented as a single bidirectional switch and the state of each converter arm is given by the binary variables S_a , S_b and S_c . S_a is defined as follows:

$S_a = 1$ when upper switch of phase A is conducting.

$S_a = 0$ when lower switch of phase A is conducting.

S_b and S_c are defined in a similar manner. Following [5], and using these logic variables, the following system equations may be written for fig. 1:

$$L_s \frac{di_1}{dt} + R_s i_1 = V_{sa} + \frac{V_c}{3} (S_b - 2S_a + S_c) \quad (3)$$

$$L_s \frac{di_2}{dt} + R_s i_2 = V_{sb} + \frac{V_c}{3} (S_c - 2S_b + S_a) \quad (4)$$

$$L_s \frac{di_3}{dt} + R_s i_3 = V_{sc} + \frac{V_c}{3} (S_a - 2S_c + S_b) \quad (5)$$

Equations 3, 4 and 5 are of the same form as equations 1 and 2 for the single phase case except that now the effective dc voltage seen by each phase is a multiple of the capacitor voltage, which (the multiple) depends on the switch states in all three phases. Note that at any given time only one switch in each converter arm conducts. It can be seen that the effective dc voltages seen by each phase take the values $\pm(2/3)V_c$, $\pm(1/3)V_c$ and 0, depending on the switch state. The zero state (when all switches in the same half of the converter are either fully on or off) gives no control of the line currents and does sometimes cause the line currents to exceed the

hysteresis boundaries. It can be shown in fact that a line current can reach a maximum of twice the hysteresis band size [2]. This is one of the limitations of the HCC scheme, but because the zero state exists only for a short period of time, this condition is negligible as will be seen from the simulation results.

Successful operation of the rectifier requires that the six antiparallel diodes remain reverse-biased at all times. This is essential in giving the device its ability to force the line currents in a desired direction, and places a lower limit on the dc link voltage. A simple analysis of the six diode stack reveals that this minimum voltage is $V_c = \sqrt{6} V_s$, where V_s = RMS phase voltage.

The factors that determine the switching frequency of the converter may be determined using equation (3) for phase A of the supply. If the effective dc voltage seen by the phase is denoted by V_a , the time (dt) in which the line current will increase by di , may be approximated by: (neglecting source resistance)

$$\Delta t = \frac{L_s \Delta i_1}{V_{sa} - V_a} \quad (6)$$

Equation (6) shows that the converter switching frequency ($\approx 1/\Delta t$) is influenced by the line inductance, supply voltage, current ripple (hysteresis band size) and most importantly, by the effective dc voltage seen (V_a). This voltage varies as the load on the rectifier changes so that the switching frequency varies with load conditions. Furthermore, since voltage V_{sa} varies over one source cycle, the switching frequency too varies over the source cycle. This variable switching frequency is the most important limitation of the HCC scheme. However, for reasonable hysteresis band sizes, the average and instantaneous switching frequencies are well within the capabilities of modern power switches such as the IGBT and Power Transistor.

4. CLOSED LOOP CONTROL OF THREE PHASE HCC RECTIFIER

As for the single phase case, the line currents for each phase are controlled according to the HCC scheme to track sinusoidal reference waveforms. Each reference waveform is generated to be in phase with the appropriate source voltage so as to achieve unity power factor. The reference current waveforms are then subtracted from the actual line currents and the differences are fed into hysteresis comparators which provide the switching signals for the three phases. A schematic representation of the current controller is

shown in figure 5. The current reference magnitude is

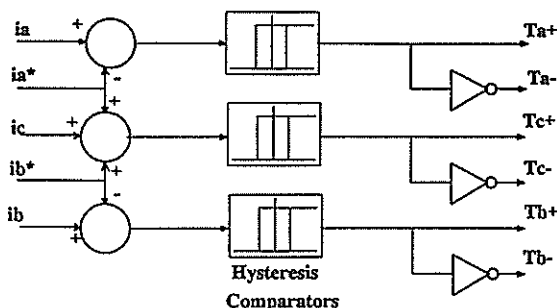


Figure 5: Current Control in Three Phase Rectifier

obtained from the output of a closed loop dc link voltage regulator as shown in figure 6. The reference magnitude is then multiplied by unity amplitude sinusoidal template waveforms for each phase to yield sinusoidal reference signals that are in phase with the appropriate source voltages. The unity amplitude template waveforms are obtained from the outputs of transformers whose primaries are connected to the phase voltage supply points as shown in figure 6. In designing the closed loop voltage regulator, only the dynamics of the dc load side need be considered since

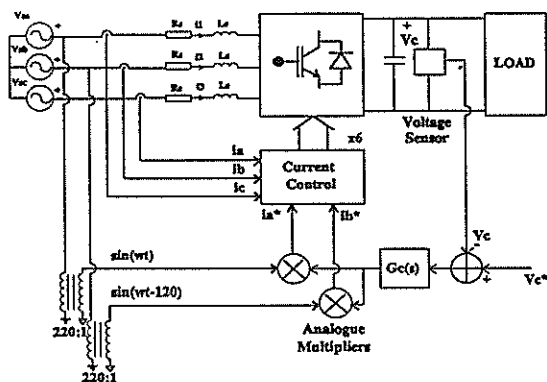


Figure 6: Closed Loop Control of Three Phase HCC Rectifier

the response of the ac side is nearly instantaneous to step changes in power. The dynamic model of the converter may be obtained from the dynamic power balance equation of the rectifier by assuming that the input ac power equals the dc link power. The model

so obtained is nonlinear and may be linearised about the operating point. The design of the dc link voltage regulator may then proceed using this simplified model. The procedure will not be discussed further here due to space limitations.

In an ac variable speed drive, regenerative operation of the ac motor would correspond with an increasing capacitor voltage. From fig 6 it can be seen that if the capacitor voltage increases to a value greater than the reference value, the reference current magnitude becomes negative. The current controllers then force the line currents to be 180 degrees out of phase with the supply voltages and power is returned from the dc link back into the ac supply. Operation during regenerative conditions is identical to operation as an inverter.

5. SIMULATION RESULTS

The HCC rectifier was simulated using the CASED package which was developed at the University of the Witwatersrand [6]. Figures 7 and 8 show the simulated converter line currents when operating as a rectifier and an inverter respectively. The near sinusoidal line currents are clearly evident. Furthermore, it can be seen that the line currents are in phase with the source voltages thus providing unity power factor. Figure 9 shows the capacitor voltage for rectifier operation. The low ripple content of the capacitor voltage is clearly evident. The system parameters were $V_s=310V$, $L_s=24mH$, $C=500\mu F$, $R_s=1\Omega$, $I_h=0.5A$ and $V_c^*=700V$. A PI voltage regulator with $K_p=0.04$ and $K_i=10$ was used.

The simulations were carried out with an R-L load in series with a constant voltage source. The load parameters were $R_o=100\Omega$, $L_o=5mH$ and $E_o=0$ for rectifier operation and $E_o=1500V$ for inverter operation.

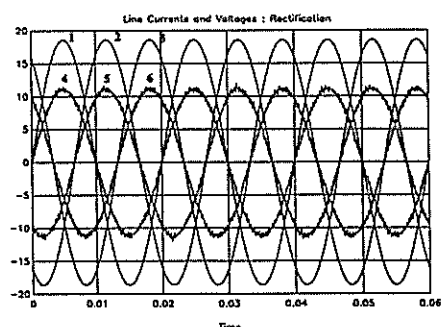


Figure 7: Line Currents and normalised Phase Voltages for Rectifier Operation

1: A phase voltage, 2: B phase voltage, 3: C phase voltage, 4: A phase line current, 5: B phase line current, 6: C phase line current

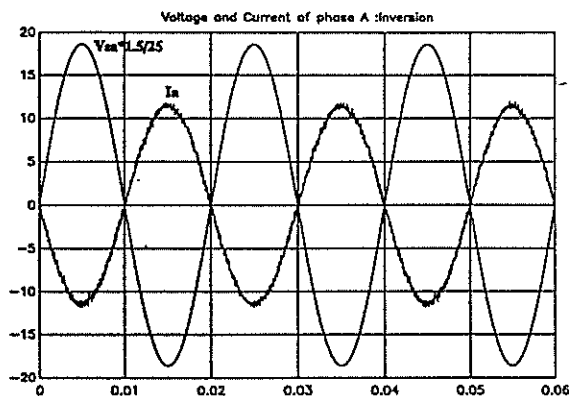


Figure 8: Line Current and normalised phase voltage : phase A (inversion)

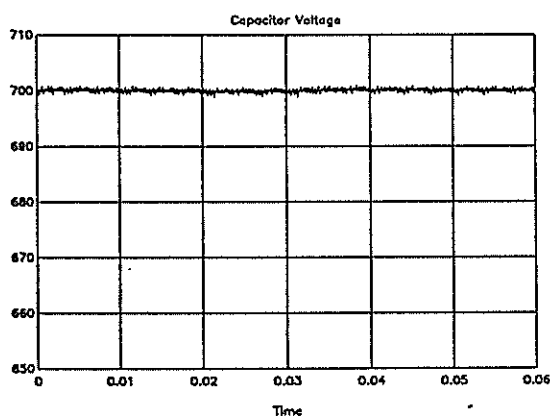


Figure 9: Steady State capacitor voltage : Rectification ($V_c^* = 700$ V)

6. CONCLUSION

This paper has described a force-commutated rectifier which is capable of achieving near sinusoidal line currents, unity input power factor and bidirectional power flow when operating under hysteresis current control. It has been seen that the device is capable of excellent performance but with a variable switching frequency.

The work discussed in this paper was done as part of the first phase of a postgraduate research project which has set out to investigate the problem of rectifier control to achieve sinusoidal line currents and unity

input power factor. The next phase of the project, which is currently being considered, is the implementation on microprocessor based hardware of a space-vector based rectifier control scheme. The space vector method of current control is characterised by a fixed switching frequency. The final phase of the project will involve practical implementation of the two control schemes after which comparisons of performance will be made from experimentally determined results.

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A NOVEL SCHEME FOR REDUCING THE HARMONICS ASSOCIATED WITH BRIDGE RECTIFIERS

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This paper describes a novel scheme for reducing the harmonics associated with bridge rectifiers. The method is based on harmonic current injection and can be applied to both controlled and uncontrolled bridge rectifiers. The background theory of the scheme is introduced along with the proposed circuit and test results.

1 INTRODUCTION

In order to limit harmonic pollution in power systems to acceptable levels, the normal approach is to install tuned filters, one per phase for each harmonic, as shown in Fig 1. The cost of such filtering schemes is high, since the filter components must be rated for high voltages and have low tolerances in order to obtain sharp tuning. Costs of filter banks can run into hundreds of thousands of dollars and companies such as ABB specialise in their design and manufacture.

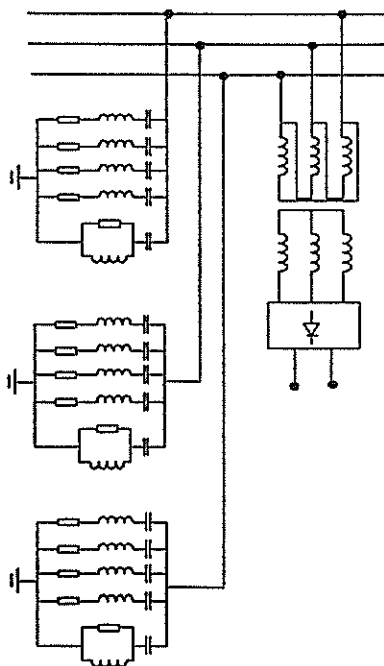


Fig. 1 Conventional passive filtering circuit

An alternative approach, in the case of transformer fed bridge rectifiers, is to inject third harmonic current into the star point of the secondary winding. This is the method proposed by Ametani in the early seventies

[1], [2]. In order to overcome the limitations of Ametani's method, namely the need for a synchronised controllable third harmonic current source, a new method of harmonic injection has been developed. This method was first described at the 1991 Australian Universities Power Engineering Conference [3] and subsequently published in the IEEE Transactions [4].

In this paper the general theory of harmonic injection is described, followed by the principle of the new scheme as applied to both a diode bridge rectifier (uncontrolled) and an SCR bridge rectifier (controlled).

2. GENERAL THEORY OF HARMONIC CURRENT INJECTION

An injected harmonic current of three times line frequency is applied to a star connected transformer secondary feeding a three phase diode bridge rectifier, as shown in Fig. 2.

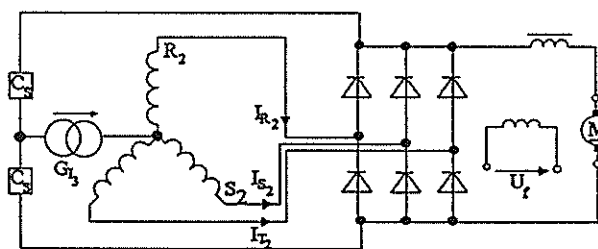


Fig. 2 Basic harmonic injection circuit

Assuming smooth dc load current, the injected current will be superimposed on rectangular blocks of secondary winding current. Further modification to the current waveshape takes place when the current is reflected across to the primary windings of the transformer. The current waveforms for the system are shown in Fig. 3.

There are two key parameters associated with the injection process

- (i) amplitude factor, $\rho = I_j / I_d$

where I_j = rms value of injected third harmonic

I_d = value of dc load current

(ii) phase shift, β

where, β is the phase shift between the peak point of the third harmonic current and the centre of the dc current pulse

It can be shown that for optimum compensation, ie minimum harmonic distortion in the line current, the value of β should equal 0° and the value of ρ should equal 0.52. [4].

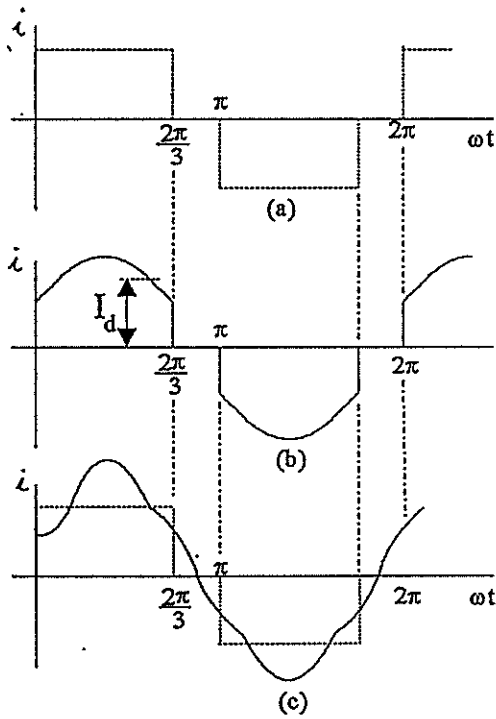


Fig. 3 Transformer currents (a) dc side- no injection (b) dc side after injection (c) ac side after injection

3. PRINCIPLE OF THE NEW METHOD

In the new scheme, shown in Fig. 4, tuned filters are connected between the rectifier output rails and the neutral point of the transformer secondary winding. The filters are designed to introduce zero phase shift in the third harmonic currents. The two filter outputs are combined at the neutral point and current is injected into the dc windings according to the conducting diodes.

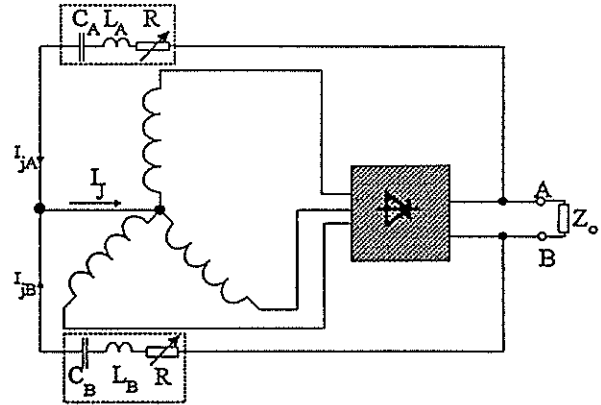


Fig. 4 Proposed circuit with tuned feedback filters

Consider the expressions for the output voltage of a three phase bridge rectifier (assume nomenclature as shown in Fig 4). The voltage between point A and the transformer neutral is given by,

$$v_{AO}(t) = k_1 V \left[1 + \frac{1}{4} \sin 3\alpha - \frac{2}{35} \sin 6\alpha + \frac{1}{40} \sin 9\alpha \dots \right]$$

where, V is the effective value of the transformer secondary phase voltage.

The voltage between point B and the transformer neutral is similar to v_{AO} , but phase shifted 60° ,

$$v_{BO}(t) = k_1 V \left[1 + \frac{1}{4} \sin 3(\alpha + 60) - \frac{2}{35} \sin 6(\alpha + 60) + \dots \right]$$

Simplifying this equation gives,

$$v_{BO}(t) = -k_1 V \left[1 - \frac{1}{4} \sin 3\alpha - \frac{2}{35} \sin 6\alpha - \frac{1}{40} \sin 9\alpha \dots \right]$$

Extracting the third harmonic components from the equations for v_{AO} and v_{BO} gives,

$$v_{AO3}(t) = k_2 \sin 3\alpha$$

$$v_{BO3}(t) = k_2 \sin 3\alpha$$

When voltage v_{AB} is considered as the difference between v_{AO} and v_{BO} then the third harmonic components cancel, however the third harmonic components in v_{AO} and v_{BO} can be used to produce harmonic currents in the transformer secondary windings. The filter components, L and C are chosen so as to ensure resonance at the third harmonic frequency. They present a low impedance to the third harmonic current and a high impedance to the 6th and

higher harmonics. The two filter currents i_{3AO} and i_{3BO} , flow independently through the upper and lower diode groups in the bridge rectifier, hence the positive and negative current pulses in the transformer windings will be modified as shown. Ideally the magnitude of the third harmonic current should be adjusted with changes in load to ensure that the amplitude factor, ρ , remains at the optimum value of 0.52.

4. APPLICATION OF THE NEW METHOD TO A CONTROLLED BRIDGE RECTIFIER

In a controlled three phase bridge rectifier the introduction of delayed firing angles will change the shape of the voltage waveforms on the rectifier output rails ie. the harmonic forcing function is changed. If the bridge circuit includes harmonic feedback filters tuned to the n 'th harmonic, the harmonic current will be modified in both amplitude and phase by the changes to the forcing function. Given that the effectiveness of the harmonic compensation is dependent upon these two characteristics of the feedback current it is apparent that the basic circuit will need to be modified for the case of a controlled bridge. The nature of the modifications will depend upon the magnitude and phase of the changes introduced by firing angle delays, hence we need to quantify the changes to the harmonic forcing function as a function of firing delay angle. The equations derived for the Fourier coefficients of the third harmonic forcing function as a function of bridge delay angle are,

$$a_3 = \frac{3}{2\pi} \left[\frac{1}{2} \sin(2\pi + 4\alpha) \sin \frac{4\pi}{3} - \sin(\pi + 2\alpha) \sin \frac{2\pi}{3} \right]$$

$$b_3 = \frac{3}{2\pi} \left[\cos(\pi + 2\alpha) \sin \frac{2\pi}{3} - \frac{1}{2} \cos(2\pi + 4\alpha) \sin \frac{4\pi}{3} \right]$$

where α is the bridge delay angle.

The above results can be used to plot the magnitude and phase of the forcing function versus bridge delay angle.

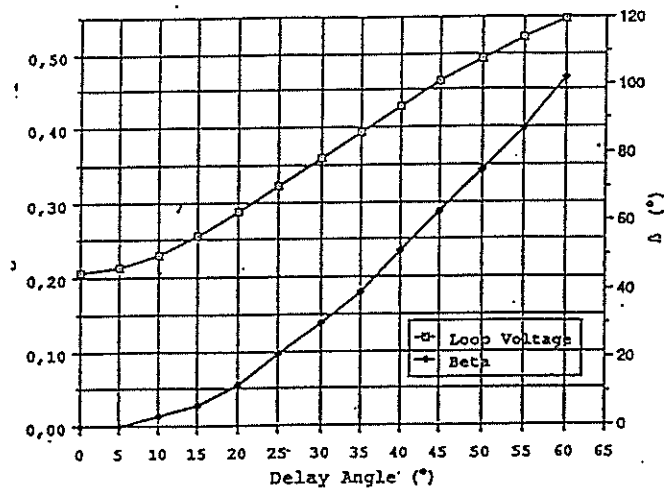


Fig 5 Per unit magnitude and phase shift of forcing function versus bridge delay angle

From the above plot several conclusions can be drawn:

(i) The magnitude of the third harmonic forcing function increases with bridge delay angle and since increases in delay angle will lead to decreases in load current, the amplitude factor will increase. (recall that

$$\rho = I_{3rms} / I_{dc})$$

(ii) The phase shift angle β increases at a faster rate than the bridge delay angle α . Hence there will be a departure from the $\beta = 0^\circ$ optimum conditions. The scheme is particularly sensitive to changes in β .

Both of these outcomes mean that if the scheme is to be effective in the case of a controlled bridge rectifier, the compensation circuit must adapt to changes in bridge delay angle and changes in load current.

The solution which is currently being investigated is to include TCRs (thyristor controlled reactors) in the feedback loops. As the bridge delay angle increases, so the current in the feedback loop must be delayed, by switching the TCR, in order to maintain the phase shift angle β equal to zero. The delay introduced by the TCRs will also attenuate the feedback current and help to restore the amplitude factor ρ to its optimum value.

5. TEST RESULTS

The scheme has been extensively tested for the case of a diode bridge and under optimum conditions the total harmonic distortion has been reduced from 24.5% down to 5.3%. Typical waveforms for line current and the associated spectrum are shown in Figs. 6 and 7.

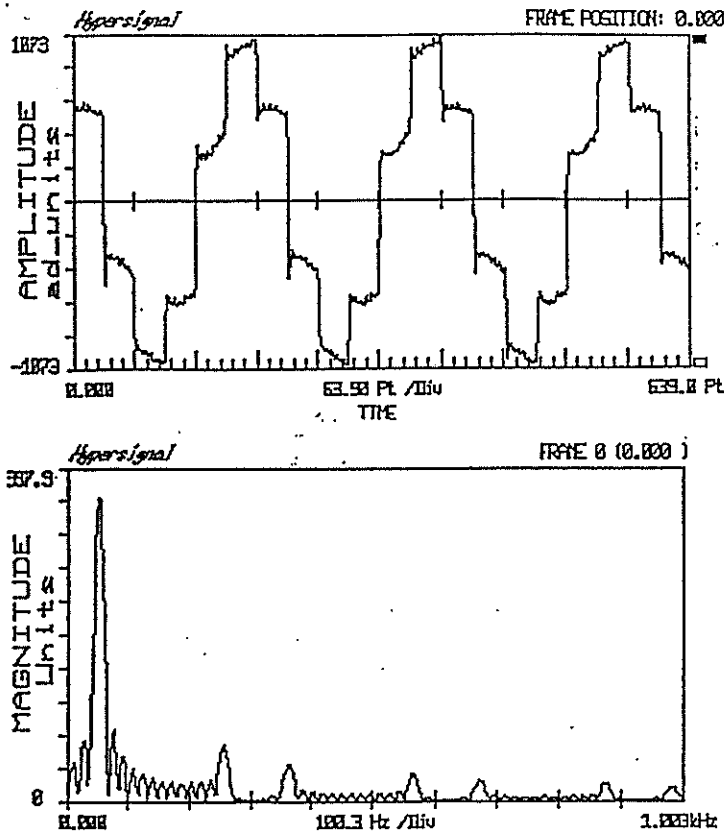


Fig 6 Line current and spectrum for a normal diode bridge rectifier, (R-L load)

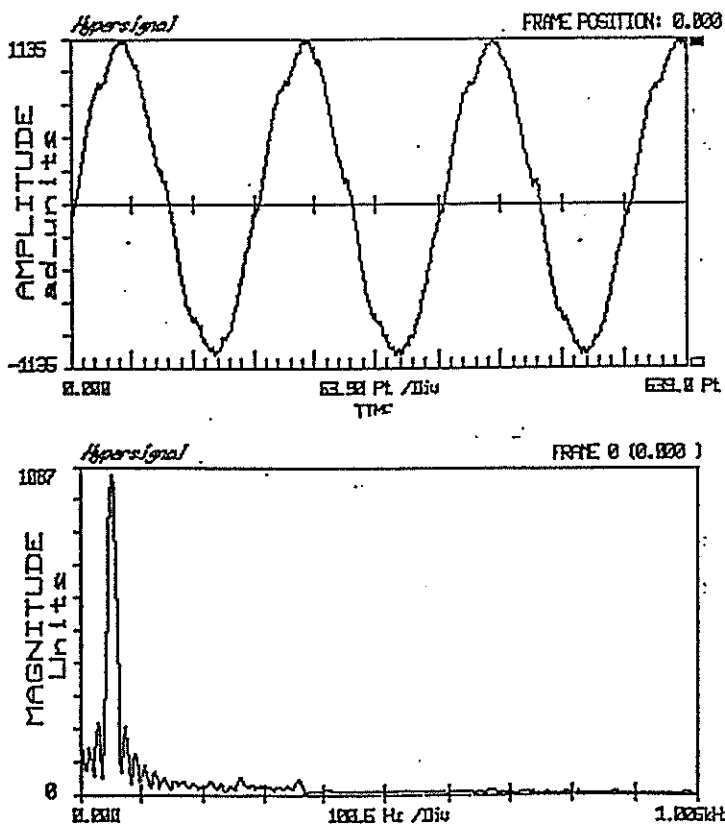


Fig 7 Line current and spectrum for a diode bridge rectifier with injection, (R-L load)

6. CONCLUSION

The new harmonic injection compensation method described, leads to a significant improvement in line current distortion associated with bridge rectifiers. Compared to conventional filters the scheme is much cheaper to implement since only two filters on the low voltage side of the system are required. Further development of the scheme is being carried out in order to apply it to a controlled bridge rectifier

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New Topologies for Voltage and Current Source Static VAR and Active Power Filters Using Soft Resonant Switching Techniques

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Abstract

New types of hard switched Static VAR Compensators(SVC's) and Active Power Filters (APF's) have response times that exceed that possible by conventional thyristor controlled inductors or thyristor switched capacitors. Limitations in switching speed of high power units does however, still place an upper limit on the response time of VAR Compensators and maximum filtering frequency of Active Power Filters. This paper discusses new resonant soft switched units that can be used in the above applications. Because switching speeds are higher, as a result of the decreased switching losses, the transient response of the resonant units is better. The new units are capable of instantaneous compensation for harmonics, random noise, bad power factor and unbalanced loads. A novel resonant notch voltage source inverter based active power filter is introduced.

1. Introduction

New types of Static VAR Compensators(SVC's) and Active Power Filters(APF's) have flourished in research environments over recent years[1-6]. These units all utilise a 3-phase hard switched inverter as an interface between the supply and a *single* energy storage element. The power circuit topologies for the inverter based Static VAR Compensator (ISVC) and Active Power Filter (IAPF), are exactly the same. Both units utilise reactive power transfer between the supply or load, generally at fundamental frequency with an ISVC and at fundamental and harmonic frequencies with an IAPF. Both units can also perform power factor correction while the main difference between the two is the control philosophies adopted for each[12-14].

Medium power hard switched ISVC's and IAPF's are operated at low switching frequencies usually in the range 600 - 800Hz[1-3]. High power ISVC's have been reported to operate with quasi-square waveforms[15]. A number of units are placed in parallel at the multiple secondary windings of a transformer connected to the supply. By phase shifting the inverter's switching signals stepped sinusoidal output waveforms are achieved.

The introduction of soft resonant switching to inverter circuits[7-13], where either or both the voltage across a switch or the current through the switch are resonated to zero, has lent itself to ISVC's and IAPF's. Faster

switching speeds are now possible with units in the high power range prospectively having switching speeds in the order 2 - 5KHz. Medium power units in the low MVA range will probably have switching speeds up to 10KHz.

Two methods of resonant switching are described here, and both methods can be used with ISVC's and IAPF's[10-13]. Resonant pulse units rely on the least number of switching components and produce resonant pulses by switching the supply voltage directly onto the energy storage element and resonant components. Individual pulses are formed which, when they return to zero, allow the inverter switch configuration to be changed with very low losses. The pulses are guided by the inverter to the output phases. Resonant notch units require more switching components associated with the resonant circuit. A DC voltage or current is always present at the energy storage element and across the inverter. The resonant circuit and its switches resonantly force the voltage across or the current through the inverter to zero for very low loss switching.

2. Hard switched ISVC's & IAPF's

Both voltage source and current source inverter based units have been constructed[1-6]. To achieve high voltage ratings multiple parallel units are connected to the secondary windings of a transformer. These configurations are shown in figure 1.

It should be mentioned that all the units have output current as a reference. The voltage source unit of figure 1c requires that an inductance be placed between the supply and itself, in this case the transformer windings, and the voltage across the inductance is controlled to give the required current. The outputs from the units of figure 1a and 1c are PWM type currents and voltages respectively.

3. Resonant Pulse ISVC's & IAPF's

A resonant pulse unit needs the same DC energy storage component as hard switched units but also introduces smaller resonant components L_{res} and C_{res} . Figure 2 shows current and voltage source units.

The operation of the current source unit requires that the supply appear as a voltage source during each resonant pulse. Capacitors are therefore used at the input. They, along with the transformer inductance, perform the additional duty of a second order filter to

remove the pulse or switching frequency from the injected current. The equivalent circuit over one resonant pulse is shown in figure 3a. The switch 'S' represents a particular inverter switching configuration and the AC voltage supply appears as a DC source over the period of the pulse.

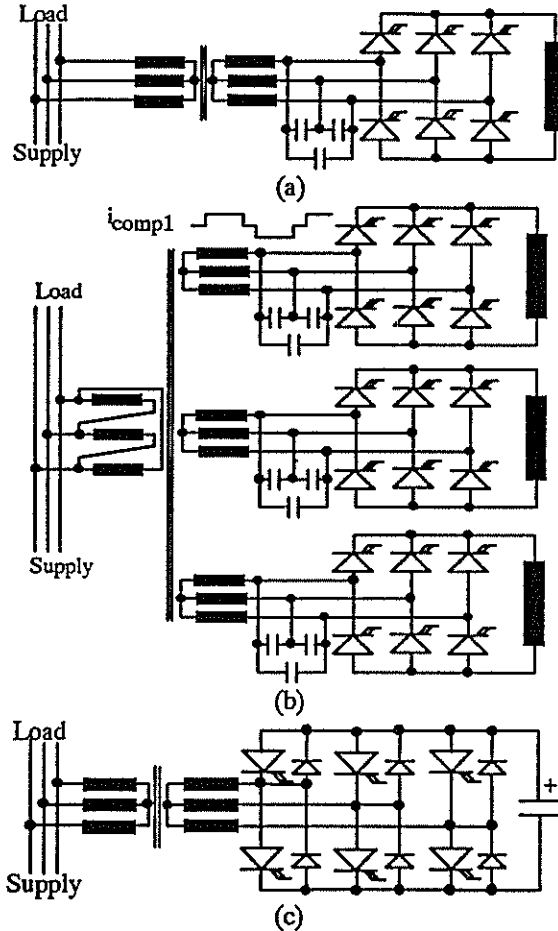


Figure 1: (a) current source ISVC & IAPF, (b) current source ISVC, (c) voltage source ISVC & IAPF. Snubber circuits are needed for all these units.

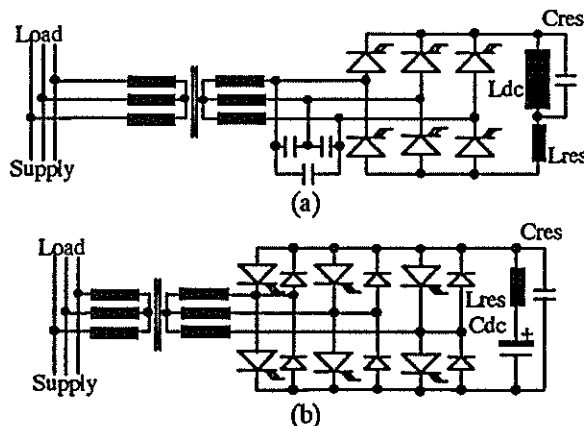


Figure 2: Resonant pulse (a) current source and (b) voltage source ISVC and IAPF.

Voltage source units need an inductance at the input so the supply appears as a current source during each resonant pulse. The voltage source circuit of figure 2b is a direct dual of figure 2a.

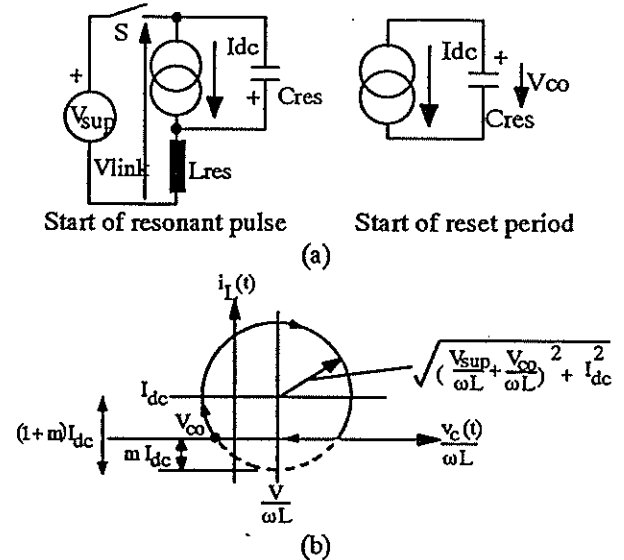


Figure 3: Resonant pulse current source unit (a) equivalent circuit during and between resonant pulses (b) current pulse i_L plotted against voltage across resonant capacitor v_c (resistive losses ignored).

Considering figure 3a, assume current I_{dc} is at steady state and C_{res} is charged negatively. Switch 'S', representing the inverter, is closed, connecting a particular line-line voltage to the circuit. The current through switch 'S' will consist of a resonant component and a DC offset, the latter due to I_{dc} . At the end of the resonant current pulse switch 'S' opens (inverter GTO's switched off with low losses) and I_{dc} automatically commutates to C_{res} and resets it to a negative voltage. Figure 3b shows the trajectory of the current pulse plotted against the voltage across C_{res} . Because the current pulses are positive, it is only necessary to use unidirectional switches such as GTO's in the inverter.

It is important that C_{res} is reset to the correct voltage (V_{co}) so that enough energy is stored in the capacitor to ensure the current returns to zero after the next pulse. Obviously, it is necessary that information about which two phase lines require a positive and negative current pulse is known during the reset period. V_{co} can be determined from equation (1) below.

$$V_{co}(\tau) \approx I_{dc} \sqrt{\frac{L_{res}}{C_{res}} (2m + m^2)} - V_{link}(\tau^+) \quad (1)$$

where m is a safety ratio as shown in figure 3b. Values of $m = 0.4 - 0.6$ have proved successful.

Figure 4 shows experimental results from a equivalent circuit model. As can be seen there are times when no pulse is required by the supply. The resonant circuit however, must continue to generate pulses so it is allowed to freewheel through the inverter.

4. Resonant Notch ISVC & IAPF

Figures 5a and c show a resonant notch current source ISVC/IAPF. Instead of injecting pulses of current as in the last case, the current injected is DC. The resonant circuit provides an alternative route for the current

during the period when the inverter switch status is required to change.

Figure 5c is a new configuration and a *dual* of the circuit shown in figure 5a. Its modes of operation are described below. The transformer, or in the case of an inverter connected directly to the supply grid, an inductor, causes the supply to appear as a current source over the short period of the notch. The equivalent circuit is shown in figures 6 and 7.

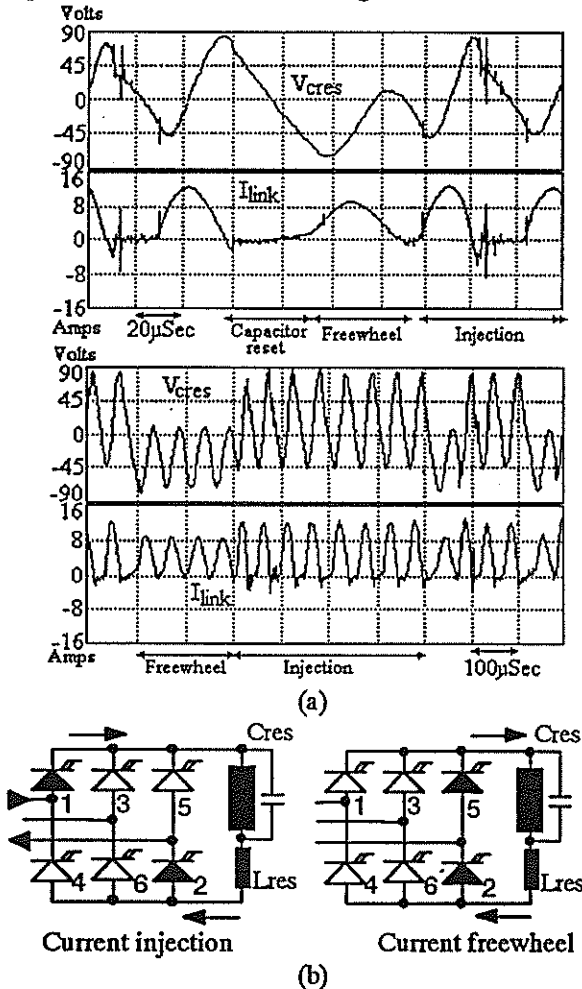


Figure 4: Current source unit showing (a) experimental results of resonant capacitor voltage and link current from an equivalent circuit model and, (b) current pulse path during injection and freewheel modes.

Referring to figures 6 & 7, Mode 1 is the normal operating condition before a notch occurs. I_{in} is either charging C_{dc} or discharging it through T1 or D1. T2 is on but not conducting. As C_{dc} is large the voltage across it, also applied to the input terminals, remains nearly constant at V_{dc} . A DC current is also flowing around the loop consisting of L_{res} -D2-T1. L_{res} must be low resistance but large enough to maintain this current between notches.

A notch is inserted into the terminal voltage so that the inverter switches can be reconfigured under zero voltage conditions. To start the notch (Mode 2), T1 is switched off. C_{res} starts charging negatively from zero

volts due to the current in L_{res} . The voltage appearing at the input terminals is given by equation (2).

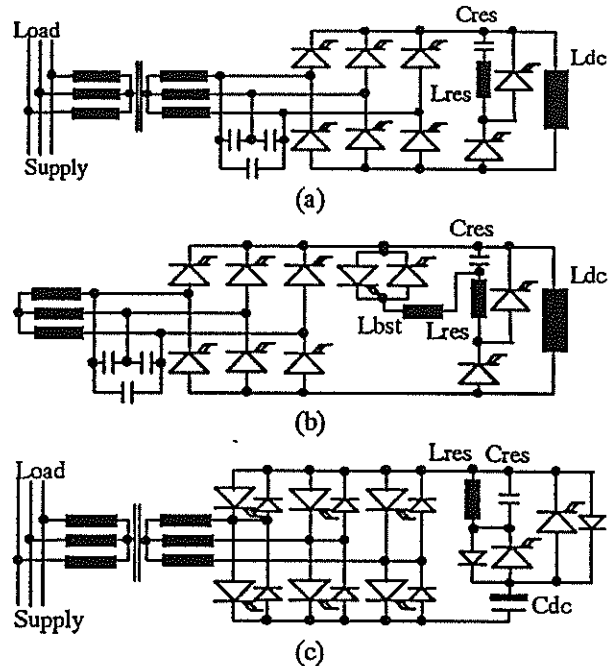


Figure 5: (a) Current source resonant notch, (b) current source resonant notch with boost circuit and (c) voltage source resonant notch, ISVC and IAPF.

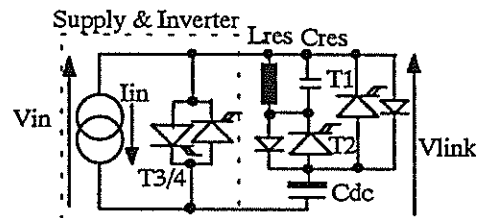


Figure 6: Voltage source resonant notch Compensator equivalent circuit.

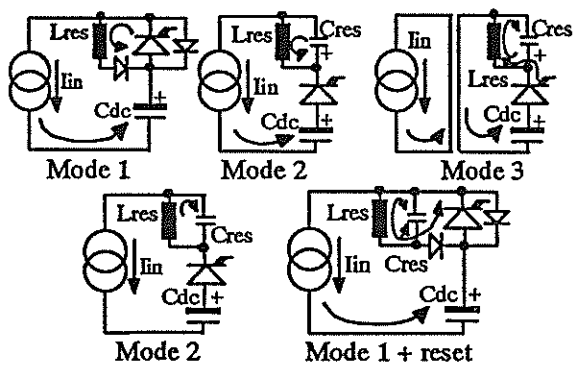


Figure 7: Operating modes for voltage source resonant notch Compensator

$$v_{link}(t) = V_{dc} - v_c(t) = V_{dc} - \sqrt{\frac{L_{res}}{C_{res}}} (I_{in} - I_{Lo}) \sin \omega t \quad (2)$$

where $\omega = 1/\sqrt{LC}$, I_{Lo} = current in L_{res} at $t = 0$.

It can be seen from equation (2) that it is important that I_{Lo} is correctly chosen so that $v_{link}(t)$ reduces to zero for the change in inverter switch configuration.

That is:

$$I_{Lo}(\tau) \approx \sqrt{\frac{L}{C}} \cdot V_{dc} - I_{in}(\tau^+) \quad (3)$$

When $v_{in}(t)$ reaches zero, T3/4 are turned on (Mode 3). C_{res} maintains $-V_{dc}$ across it and the current in L_{res} increases linearly. When this current reaches I_{Lo} , T3/4 is switched off (Mode 2) and the current in L_{res} resonantly charges C_{res} in a positive direction. As the voltage across C_{res} goes through zero ($V_{in}(t) = V_{dc}$), T1 is switched on and T2 off (Mode 1 + reset). V_{dc} is now reconnected directly to the input terminals.

C_{res} continues to charge in a positive direction and once fully charged will resonantly discharge through L_{res} . As the voltage across C_{res} reaches zero and just starts going negative, diode D2 conducts and the full I_{Lo} will now freewheel through L_{res} -D2-T1. The circuit is ready to start the next notch.

5. Results

Figure 8 show simulation results of a pulsed current source IAPF performing power factor and harmonic compensation. The sudden application of an additional unity power factor load on phase 'b' results in the compensator immediately balancing the supply current as a natural part of its control philosophy[14]. It reaches a balanced steady state after 60°(3.3mS).

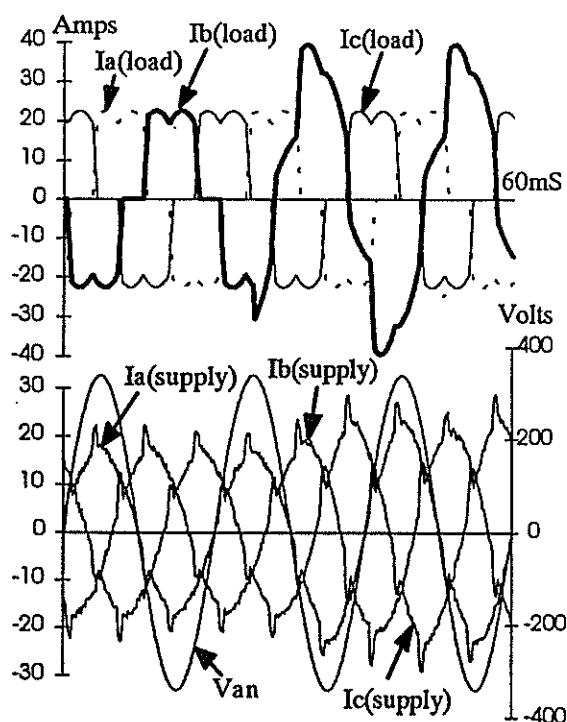


Figure 8: (top) load current drawn by 3-phase controlled rectifier with $\alpha = 40^\circ$ and resistive load suddenly applied to phase 'b', (bottom) phase 'a' voltage and supply currents after power factor and harmonic compensation. The unit is connected to the supply via a star/star transformer with the primary neutral connected.

6. Conclusion

This paper has shown a range of possible configurations for both inverter based static VAR compensators and active power filters using resonant soft switching techniques. There is a strong expectation that capital costs associated with plant ie filters, power factor correction capacitors and static switched capacitors and inductors for VAR compensation, will be reduced if ISVC/IAPF units are used. This combined unit with its *single* energy storage element, can perform all the above compensation *at the same time*. A new voltage notch unit has been introduced. Unbalanced loads are shown to be compensated by the IAPF which results in balanced supply currents using no overall steady state real power flow. Transient load changes however, are smoothed by using real power flow from the energy storage element.

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Modelling of a Capacitive Voltage Transformer for the Electromagnetic Transients Program (EMTP)

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Abstract — This paper presents a broadband lumped parameter equivalent circuit model a typical capacitive voltage transformer (cvt) using software simulation tools such as PSPICE and the Electromagnetic Transient Program (EMTP). It is shown that parasitic elements such as the winding capacitances of the intermediate voltage inductor and the intermediate voltage transformer have important effects on the frequency response of practical cvts. The role of the transient damping arrangement and the secondary burden is discussed with reference to the Q-factors of the resonance peaks associated with the overall frequency response.

1. INTRODUCTION

The transient response of a cvt exhibit resonant characteristics for disturbances introduced at the high or low voltage terminals. This resonant behaviour has two important practical effects, namely a significant degree of distortion of the output signals and increased insulation stresses due to possible overvoltages at internal nodes. Although a typical cvt features a transient damping arrangement to suppress such overvoltages, insulation failures due to overvoltages still occurs in practice. This can be readily investigated using simulation programs such as PSPICE and EMTP, provided suitable model structures and parameter data are available. Although some experimental results with reference to the frequency domain behaviour of cvt's have been published [3, 6, 8], information on the modelling of cvt's for broadband and transient applications are limited. A good equivalent circuit have been proposed by Kezunovic *et al* [5, 6], and frequency response and transient response results have been presented. This paper presents an equivalent circuit model which essentially confirm the validity of the circuit proposed by Kezunovic *et al* [5, 6], and considers additional complexities which may arise in the modelling of the intermediate voltage inductor. Measured parameter data are given for a commercial 400 kV cvt. Figure 1 shows a circuit diagram of the 400 kV cvt used for the purpose of this investigation. The following definitions apply:

- C₁ High voltage capacitor
- C₂ Intermediate voltage capacitor
- C_{d1}, C_{d2} Complemental high voltage capacitors
- G Protective spark gap

- L Intermediate voltage inductor
- A Lightning arrestor
- L_c Carrier frequency drain coil
- W_c Calibration windings
- W_s Secondary windings
- L_{d1}, L_{d2} Damping circuit inductors
- C_d Damping circuit capacitor
- R_d Damping circuit voltage dependent resistor

Only the secondary winding rated for 110√3 V will be considered.

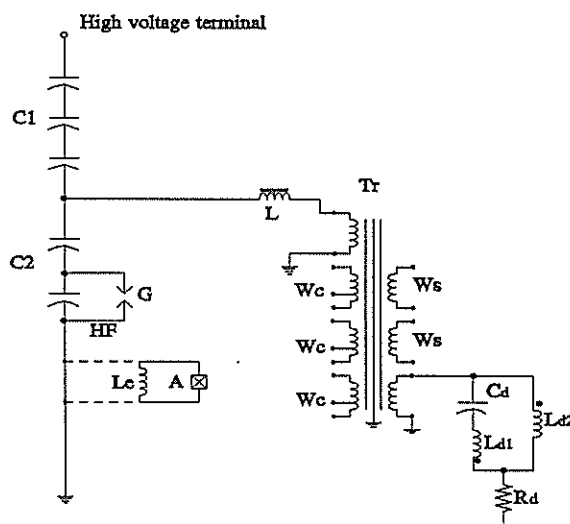


Figure 1: Circuit diagram of the test cvt.

2. LUMPED PARAMETER EQUIVALENT CIRCUIT REPRESENTATION

The equivalent circuit of the cvt will be developed with the view that it is required to represent the frequency domain characteristics of the actual cvt for the frequency range up to 2.5 kHz. This ensures a valid model for simulation studies involving harmonic distortion and transient fault waveforms. It is convenient to derive the overall equivalent circuit with reference to the various subcircuits, namely the capacitive divider, intermediate voltage inductor, intermediate voltage transformer, transient damping circuit and the burden. The overall frequency response was determined for a sealed unit, while the experimental data applicable to the various subsections were obtained for an identical unsealed unit.

2.1 Capacitive divider

The high voltage capacitors comprising the capacitive divider, namely C_{d1} , C_{d2} , C_1 and C_2 can be reduced to yield an equivalent capacitor C_1' , where

$$C_1' = \frac{C_{d1}C_{d2}C_1}{C_{d1}C_{d2} + C_{d1}C_1 + C_{d2}C_1} \quad (1)$$

The high voltage source and the divider formed by C_1' and C_2 can be reduced to the Thevenin equivalent circuit shown in figure 2, where

$$C_{Th} = C_1' + C_2 \quad (2)$$

The values for C_{d1} and C_{d2} can be obtained from the nameplates of the units, while the values for C_1 and C_2 were determined experimentally for the unsealed unit. Table I gives the value calculated for C_{Th} .

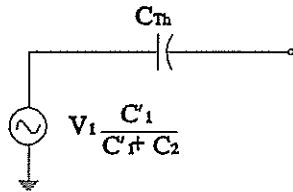


Figure 2: Lumped parameter equivalent circuit representation for the capacitive divider.

2.2 Intermediate voltage inductor

The intermediate voltage inductor consists of a series connection of two coils located on parallel legs of a common iron core. The coupling point between the coils is connected to the core. Due to the physical separation of the coils a coupling factor significantly less than unity is required. The coils have significant winding capacitances but interwinding capacitance is small due to the distance between the coils. Winding resistances must be allowed for due to the large number of turns and small conductor diameter involved. The above considerations give rise to the fourth-order equivalent circuit shown in figure 3, where the following definitions apply:

R_{L1}, R_{L2}	Winding resistances of the first and second coils respectively (Ω).
L_{L1}, L_{L2}	Inductances of the first and second coils respectively (H).
C_{L1}, C_{L2}	Winding capacitances of the first and second coils respectively (F).
M_L	Mutual inductance between the first and second coils (H).

For identical coils, as is the case for the test specimen,

$R_{L1} = R_{L2} = R_L$, $L_{L1} = L_{L2} = L_L$ and $C_{L1} = C_{L2} = C_L$. The inductor can then be represented a second-order model as shown in figure 6, where

$$R_L' = 2R_L$$

$$L_L' = 2(L_L + M_L)$$

and

$$C_L' = \frac{C_L}{2}.$$

Table I gives the parameter values determined experimentally for the inductor of the unsealed unit.

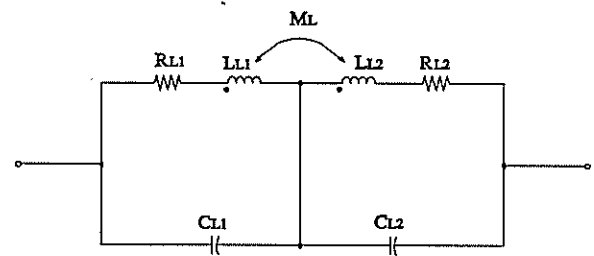


Figure 3: Lumped parameter equivalent circuit for the intermediate voltage inductor.

2.3 Intermediate voltage transformer

The intermediate voltage transformer can be modelled by means of the well-known lumped parameter equivalent circuit shown in figure 4. The following definitions apply (all parameters referred to the high voltage winding):

$R_{T11}, R_{T21}, R_{Td1}$	Winding resistances of the high voltage, low voltage and damping windings (Ω).
R_{Te1}	Core loss resistance (Ω).
$L_{T11}, L_{T21}, L_{Td1}$	Leakage inductance of the high voltage, low voltage and damping windings (H).
L_{Te1}	Magnetizing inductance (H).
C_{T11}	Winding capacitance of the high voltage winding (F).

The interwinding capacitances and the capacitances of the low voltage windings only affect the frequency response of the circuit at frequencies well above 20 kHz and have been neglected. Table I presents parameter values determined experimentally, using a combination of the techniques proposed by Douglass [4], Bak-Jensen [1] and Vermeulen [9].

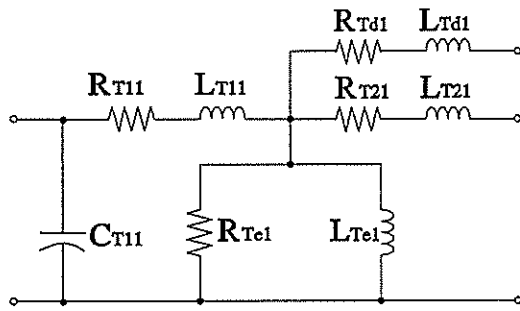


Figure 4: Lumped parameter equivalent circuit for the intermediate voltage transformer.

2.4 Transient damping circuit

The test cvt employs a passive damping circuit comprised of an LC-filter in series with a nonlinear resistance. Figure 5 presents a third-order equivalent circuit model for the damping circuit. The filter section is comprised of L_{dL1} , R_{dL1} , L_{dL2} , R_{dL2} and C_d . Table I gives measured parameter values for the filter section.

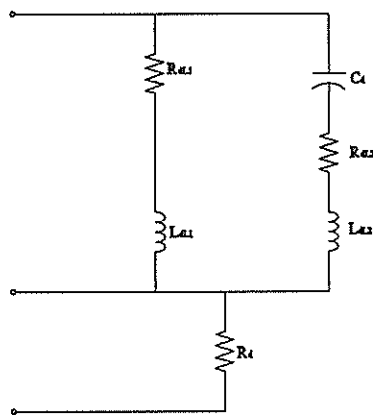


Figure 5: Lumped parameter equivalent circuit for the transient damping arrangement.

2.6 Secondary burden

Burden impedances usually assume the form of a series resistive-inductive arrangement as suggested by BS 3941:1975 [2], which specifies the standard values of rated output for a lagging power factor of 0.8. If applied to the test specimen, this yields the values given in table I for R_{B1} and L_{B1} , where R_{B1} and L_{B1} denote the resistive and inductive components respectively of the burden impedance referred to the high voltage side.

3 EQUIVALENT CIRCUIT FOR THE COMPLETE CVT

The equivalent circuit of the overall cvt comprises

an interconnection of the various subcircuits. This yields the circuit shown in figure 6. The frequency response of the circuit shows a resonant peak at 200 Hz due to series resonance of C_{Th} , L_L and C_{T11} . The response has a dip in the vicinity of 1585 Hz as a result of parallel resonance of the subcircuit comprised of R_L' , L_L and C_L' . The inductor presents a very high series impedance in the vicinity of this frequency. If viewed in conjunction with the relatively lower input impedance reflected by the intermediate voltage transformer, the resultant divider effect cause severe attenuation of such frequencies in the transfer characteristic of the overall cvt. The damping circuit does not affect the response in the vicinity of the rated operating frequency of 50 Hz. This is due to the high impedance associated with parallel resonance of L_{dL1} , L_{dL2} and C_d , which effectively decouples the damping resistor from the rest of the circuit at this frequencies. Series resonance of the damping circuit filter section however gives rise to a low value of impedance at approximately 950 Hz, so that a considerable degree of damping is introduced by the damping resistor at this frequency. The burden increases the model order of the overall arrangement and must therefore be modelled as an integral part of the system. It mainly affects the damping factors at the various resonant frequencies.

TABLE I
PARAMETER VALUES FOR THE EQUIVALENT CIRCUIT
IN FIGURE 6.

Para-meter	Value	Para-meter	Value
C_{Th} (nF)	57.18	R_L' (Ω)	766
L_L' (H)	167.6	C_L' (pF)	60
R_{T11} (k Ω)	1.416	R_{T21} (H)	2.244
R_{Td1} (k Ω)	5.184	L_{T11} (H)	2.84
L_{T21} (H)	2.84	L_{Td1} (H)	3.70
R_{Te1} (M Ω)	10.6	L_{Te1} (kH)	49
C_{T11} (pF)	276	R_{dL1} (Ω)	4.2
L_{dL1} (mH)	813	R_{dL2} (Ω)	0.37
L_{dL2} (mH)	8	C_d (uF)	10
R_{B1} (k Ω)	645.2	L_{B1} (kH)	1.54

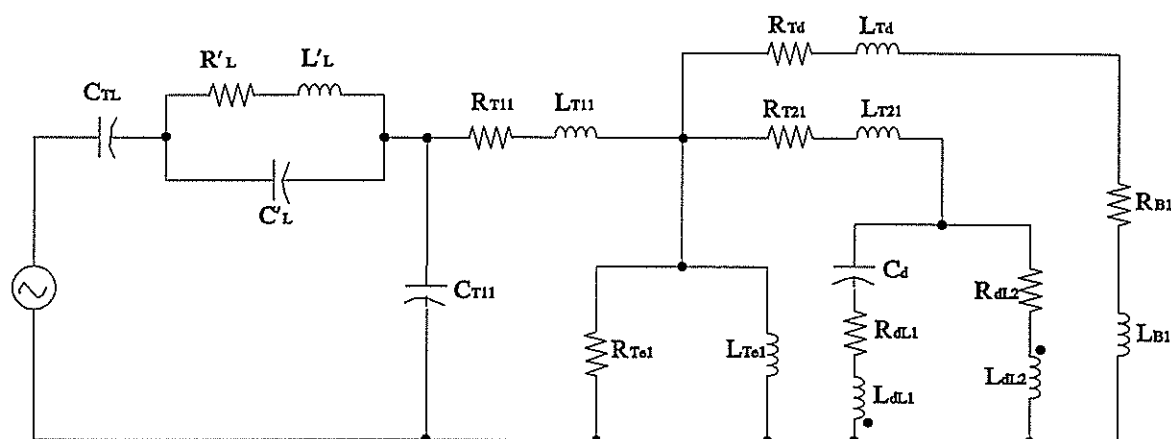


Figure 6: Equivalent circuit for the complete capacitive voltage transformer.

4 CONCLUSION

The frequency response of practical cvt's exhibit marked resonant characteristics in the region of the lower order harmonic frequencies due to parasitic circuit elements such as the winding capacitances of the intermediate voltage inductor and the high voltage winding of the intermediate voltage transformer. With the exception of the transient damping arrangement, constructional differences between various cvt's affect the parameter values rather than the equivalent circuit topologies, so that the proposed equivalent circuit can be applied for a reasonably wide range of cvt's.

6 ACKNOWLEDGEMENT

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THE MODELLING OF ARC FURNACE FLICKER AND SVC COMPENSATION REQUIREMENTS

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ABSTRACT

The application of an EMTP model based on three-phase stochastic arc voltages is discussed for a practical 33MVA furnace. The model includes a digital implementation of the UIE flickermeter to assess the flicker severity produced by the model for comparison with field measurements. Flicker-compensation using a static-Var compensator is illustrated using the EMTP model.

1. INTRODUCTION

Arc-furnaces are the largest flicker-producing loads on a power network, and have on several cases presented a problem in the Eskom network. The flicker produced by a particular arc-furnace installation is dependent on the mode of operation (particularly severe during the initial bore-down) and on the material being processed.

The planning of these loads has been facilitated by the development of various empirical techniques for assessing the expected flicker levels at the design stage [1]. However, the success of compensation and mitigation strategies such as static Var compensators, synchronous condensers, and series reactors is difficult to assess at the design stage using these empirical formulae. A proposed formulae has recently been presented [2] based on a sample of 31 furnace/SVC installations.

In order to aid the designer when assessing various compensation options for a particular new installation, several modelling techniques have been researched. These techniques are based on: field measurements of existing furnace currents [6], single-phase weighted white-noise [7], arc-resistance, or arc-voltage models [8].

The technique based on field-measurements of furnace current suffers from the short-coming that compensation can significantly change the nature of the arc. The single-phase white-noise approximation does not allow particular shunt compensator control

algorithms to be evaluated. Both the arc-resistance and arc-voltage techniques have similar benefits. The latter is described and applied to an existing 33MVA furnace installation in this paper.

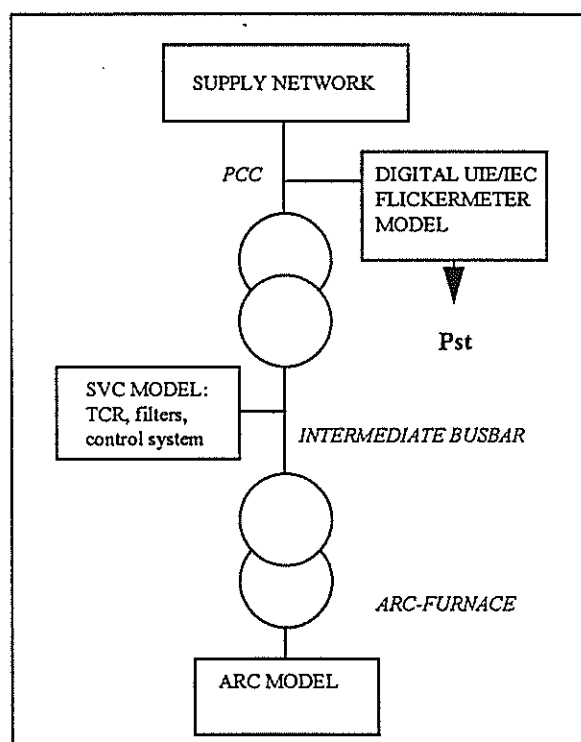


Figure 1. Overview of the arc-furnace, UIE-flickermeter, and static Var compensator simulation.

3. EMTP IMPLEMENTATION OF THE ARC-VOLTAGE MODEL

The magnitude of the arc-voltage depends principally on the length of the arc and has only weak dependence on the arc-current. As the furnace load is determined by the arc-voltage, control of the furnace is achieved by displacement of the electrodes. The time-constant limitations of this control in responding to changes in the arc length (caused by arc movement and melt collapse) result in fluctuating arc-voltages. Statistical studies [3] have shown the

Table 2. Network parameters for the modelled 33MVA furnace installation (Tap 12 - boredown).

R - total secondary	0.415 mOhm
X - furnace cable	2.649 mOhm
X - furnace transformer	0.324 mOhm
132/22kV transformer	10.5%
132kV fault level (PCC)	3546MVA

Field measurements

One of the important questions in the study was what the contribution of the plant rolling mills would be to the overall flicker levels at the PCC. Flicker measurements were therefore performed using a UIE flickermeter at the 132kV busbar VT's for a 1-week period. Furnace currents and the supply currents to the remaining plant (including rolling mill) were measured concurrently. Figure 5 shows the measured flicker severity (P_{st}) levels for a 24 hour period (red phase). Figures 3 and 4 show the corresponding furnace and mill currents (both red phase).

The measurements show that the furnace is the most significant contributor to the relatively high levels ($P_{st} = 1.3$). The mills, also a potential cause of flicker, had minimal effect on the measured 132kV flicker levels.

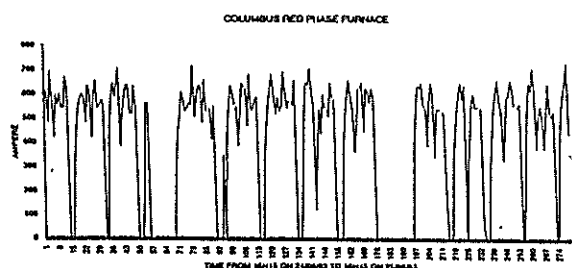


Figure 3. Measured arc-furnace currents as supplied by single dedicated 40MVA transformer..

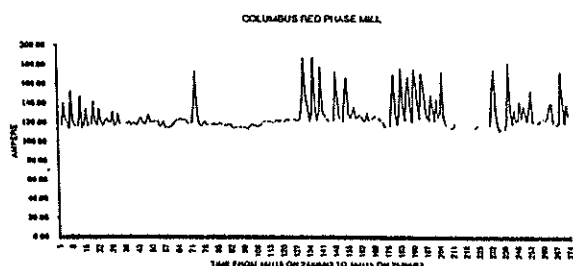


Figure 4. Measured plant currents (including rolling mills) for one of three parallel 40MVA transformers.

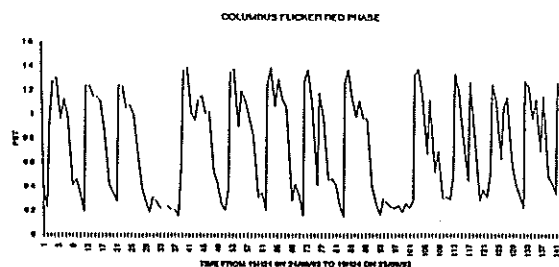


Figure 5. Measured flicker severity levels (P_{st}) on the common 132kV busbar.

Modelling

The EMTP network-, furnace-, and SVC models were implemented on a 80486DX2 micro-computer. The simulation time-step was chosen to provide the SVC model with an a 0.5 degree thyristor fire-angle delay resolution. The PCC and intermediate bus voltages were written to an output file at a rate of 800 per second.

Calibration of the model was based on the measured flicker severity values (1.3 max, 0.3 background). With the reasonable assumption that the fluctuating furnace currents were uncorrelated with the background voltage fluctuations, the contribution of the furnace to the measured flicker severity levels was calculated as ≈ 1.3 (i.e. sum of the squares).

For initial bore-down with furnace transformer Tap 12 (409.9 V @ nominal 44.59kA, 33 MVA, power factor 0.8) the modelled arc-voltage mean was chosen as $\bar{E} = 200$ V. The variance of the arc voltage was determined as $\sigma = 30$ V for a P_{st} of 1.3 at the modelled 132kV busbar.

Results

The maximum simulated P_{st} is given in Table 3.

Table 3. Predicted P_{st} values using the EMTP arc-voltage model.

	Pst : system normal
no compensation	1.36
SVC compensation	0.81

Figures 6 and 7 show the simulated arc-voltages and furnace currents.

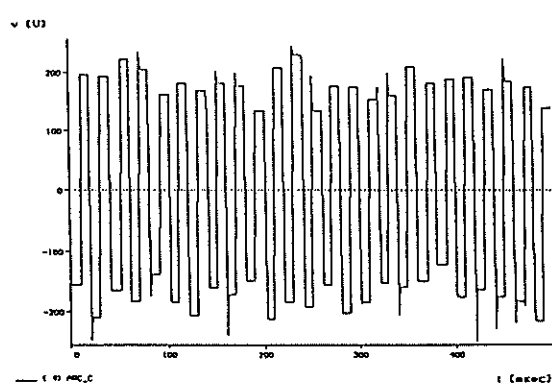


Figure 6. Simulated arc-voltages.

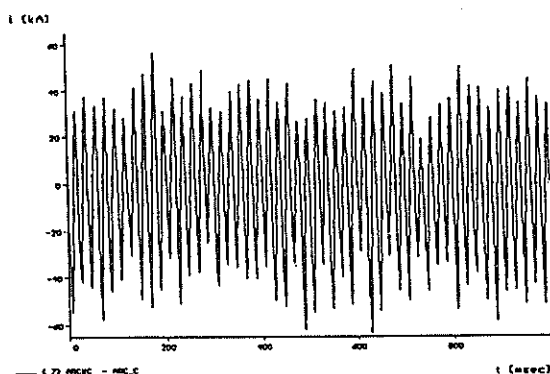


Figure 7. Simulated and measured currents.

The latest work on empirical formulations assessing the effect of shunt compensation, gives a flicker reduction factor, R_{svc} , for SVC compensation as:

$$R_{svc} \approx (1 + 0.75(S_{svc}/S_{furn})) \quad \dots(2)$$

where S_{svc} and S_{furn} are the SVC and furnace ratings respectively. For the above study, the SVC was sized as 40MVAR, giving a projected R_{svc} of 1.9. The simulated reduction was 1.68. This figure, although somewhat lower, falls well within the distribution of points from which (2) was derived [2]. Two SVC's in the above study with a similar ratio to the furnace MVA, resulted in reduction factors of 1.5 and 1.6.

7. CONCLUSIONS

- a) The applied model allows the relative effects of flicker-mitigation techniques to be assessed for a given installation (i.e. shunt compensation at the intermediate busbar, the inclusion of a series reactor, dedicated supply etc). Application of the model has been illustrated for a 33MVA furnace on the Eskom network.

- b) The model is calibrated by making use of the UIE/UNPEDE study results. Absolute flicker values can therefore only be approximated.
- c) Application of the model to several South African furnace installations must be investigated in order to gain more confidence in the results.

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To P.M. Roussow of Eskom T-R-I for measurements.

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Digital Simulation of Devices for Flexible AC Transmission Systems Using EMTP

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ABSTRACT

In this paper, a new FACTS (Flexible AC Transmission Systems) device is researched by simulation using EMTP, which can both regulate the phase angle of an injection voltage and isolate voltage harmonics between the transmission line and the load. This simulation program includes the models of the transformer, converter and load. The controlling model using the ATP scheme is adopted. In this program, more than 70 variables can be drawn by means of the PCPLOT. The program can easily be adopted to research other FACTS device characteristics.

INTRODUCTION

The Flexible AC Transmission System is akin to high-voltage dc and related thyristor developments, designed to overcome the limitations of the present mechanically controlled ac power transmission systems. By using reliable, high-speed power electronic controllers, this technology offers five improvements for increased efficiency: 1. Greater control of power so that it flows on the prescribed transmission routes; 2. Secure loading (but not overloading) of transmission lines to levels nearer their thermal limits; 3. Greater ability to transfer power between controlled areas so that the generation reserve margin (typically 18 percent) may be reduced to 15 percent or less; 4. Prevention of cascading outages by limiting the effects of faults and equipment failure; 5. Damping of power system oscillations which could damage equipment and/or limit usable transmission capacity [1], [3], and [4]. With the developing of electronic technology, larger power electronic devices are developed. Because power electronic devices deal with harmonics, a new concept about FACTS devices is proposed. With this concept the voltage harmonics between the transmission line and the nonlinear load can be isolated while regulating the impedance of the transmission line [5] and [6].

In this paper the transient performance of the phase shift controller is examined and effects of the harmonics produced by the nonlinear load is considered. Because power and control system characteristics of FACTS devices are complex, EMTP (Electromagnetic Transient Program) is

adopted to simulate the characteristics of the system.

The Electromagnetic Transients Program, or EMTP, is primarily a simulation program of the electric power industry. It can predict variables of interest within electric power networks as function of time, typically following some disturbance such as the switching of a circuit breaker or a fault. It also is used in specializing studies in power electronics, [2] and [7].

EMTP is a strong tool for power engineering problems, able to simulate not only the transient operation of rotating machinery and other power devices, but also the steady state operation. Although this program is widely used, it shows some drawbacks, such as strictly ordered and complex formatted data input.

EMTP SIMULATION PROGRAM

The simulation program is programmed in Fortran and it deals with the main EMTP models as follows:

1. TACS-controlled switch

TACS is an acronym derived from the name "Transient Analysis of Control Systems". The switch status within EMTP is controlled by TACS (Transient Analysis of Control Systems). Recall that TACS is the EMTP " digitally-simulated analog computer". Such data cards are collectively referred to as switch cards and these cards follow the regular EMTP branch cards, and precede the EMTP source cards. In this paper this model is used to simulate the thyristor rectifier, IGBT inverter, and nonlinear load.

2. TACS data cards for control system modelling

In very general terms, TACS directly provides the user with modelling that normally is associated with an analog computer. TACS data is optional for EMTP. As already mentioned, TACS functions like an analog computer. But it adds to this algebraic and logical processing, so it really is more general. It is particularly useful in EMTP because signals can be passed between the electric network and

TACS, with the result functioning like a hybrid computer. In this paper, this TACS model is used to generate control signals for the thyristors and IGBTs. The feedback signals are also handled with TACS.

3. Branch cards

Branch cards are cards where one can define different kinds of branches. Distinction should be made between linear and non-linear branches. In this paper, only linear branches are used. The branch cards can deal with uncoupled or single phase, lumped series RLC branches by specifying the non-zero impedance of a branch between two named nodes, or between one named node and ground. This impedance can be either a resistance, a capacitor, an inductor or a series connection of any of these.

All the examples presented in this paper were run on a PC-compatible computer; PCPLOT was used to plot the program responses. Because EMTP software is employed, a complex system can be simulated. Using these simulation results, a prototype FACTS controller can be developed. The simulation results are very useful to build a real system.

By means of the program, we can obtain the characteristics of the input and output sources. When there are harmonics in the input sources or when the loads are nonlinear, the harmonic effects must be considered. Eliminating the affects of the harmonics of nonlinear load is one of the purposes of this paper.

THE PRINCIPLE OF THE NEW FACTS DEVICE

According to the principle of the phase shifter, if the phase angle of the injection voltage is regulated, the power flowing through the transmission line can be changed as shown in equation (1). (To see Fig.1)

$$P = V_s \frac{V_g}{X_L} \sin \delta \quad (1)$$

According to the principle of the active filter, if the electronic equipment can provide harmonics to the transmission line, which are the same as harmonics produced by the nonlinear load, the voltage harmonics can be isolated between the transmission line and the nonlinear load [8]. The new FACTS device proposed in this paper combine the two principles together by a special control scheme.

All waveforms of the branch currents and the node voltages in the system can be obtained. They provide a convenient condition to research the variable rules of the rectifier and the inverter circuits. The simulation results can easily show the voltage across the IGBT and the action of the snubbers in the rectifier circuit and inverter. When loads are changing, the output voltage and current will be effected, on the other hand, the characteristics of the system will also be variable, so the effect of varying loads must be considered by simulation.

SIMULATION PARAMETERS

In this section, the phase shift controller and active filter with a nonlinear load is researched by simulation. The block diagram of the phase shift and active filter using the same transformer is shown as Fig.1. The power transformer 1 is Y- Δ connection. The transformer 2 is Y connection in its primary windings and its secondary windings are in series with the transmission line. The main parameters of the simulation are shown in Table I.

Table I. Main Simulation Parameters (Three Phase)

Load inductance	200 mH
Filter inductance (L_o)	3.5 mH
Filter capacitor (C_o)	500 μ F
Transformer primary leakage inductance	0.03 mH
Transformer secondary leakage inductance	0.1 mH
Series transformer turns ratio	2.5 : 28
Transmission line current	95 A

The AC-DC converter is a three phase converter, comprising six thyristors. The DC-AC inverter consists of six IGBTs.

SIMULATION RESULTS

In this paper, the DC voltage is provided by the full bridge thyristor rectifier. The functions of the phase shifter and the active filter using the same

transformer and converter are researched through simulation.

The single phase result of this simulation is plotted in Fig. 2. All the voltages and currents plotted in Fig. 2 are labelled accordingly to the labels used in Fig. 1. The characteristic distortion of the load voltage V_g (resulting from the 6 pulse thyristor converter and network impedance X_n) before the operation of the harmonic isolator, is clear from the simulation result. After the unified converter injects the voltage V_{pq} , the voltage harmonics are isolated as shown in the V_r waveform. In the process the load voltage is affected as is shown in V_g . In all the simulation cases the load current I_L is kept constant. A small phase shifter is included in the reference.

The simulation results show that this control scheme can compensate for the harmonic voltages. Using the program, we can study transient performances of the inverter and converter as shown in Fig. 1. Using PCPLOT, up to 80 variables can be plotted.

CONCLUSION

In this paper, simulating a new FACTS device concept is performed, which includes active power filter and phase shift controller using the same converter. The characteristics of the active filter and a phase shift controller with a nonlinear load is researched. The simulation results show that the method proposed by this paper both can compensate the harmonic voltages produced by the nonlinear load and can regulate the voltage of the transmission line. The leakage inductance of transformer 2 which the active filter and phase shift use to transfer energy to the transmission line, obviously affects the compensation of harmonics. So it is better to reduce the leakage inductance as small as possible. In the DC link circuit, a proper value of filter inductor L_4 will be able to reduce the amplitude of the current spikes flowing through the secondary coil of the transformer 2. A small voltage ripple in the DC link has no obvious effect on the output voltage waveform of the phase shift controller. In the simulation research, the filter capacitors are only fixed in the low voltage side so as to reduce the spikes of the voltages and the currents. The simulation results will be useful to build a practical prototype FACTS device.

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THE THERMAL MODELLING AND ANALYSIS OF UNDERGROUND CABLE SYSTEMS

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ABSTRACT - Underground cable systems are 10 - 20 times more expensive than the equivalent overhead system. It is therefore desirable to utilize the underground cable to its maximum safe working capacity without damaging the conductor or its insulation due to overheating. This paper describes preliminary experimental work done to obtain a thermal model for a XLPE cable in free air. Details are also presented of an experimental set-up which will be used to heat the cable and the surrounding soil. A data logging system has been built to store the data obtained from temperature sensors in the soil.

1. INTRODUCTION

A convenient way to deal with heat flow problems is to make use of electrical analogies. Heat flow through a thermal resistance can be calculated by using the following equation:

$$P = \frac{\Delta T}{R_t} \quad (1)$$

In this equation P is the heat generated (in Watts), ΔT is the temperature rise with respect to the ambient and R_t is the thermal resistance. Equation 1 is known as Ohm's thermal law and the following analogies with respect to electrical quantities are identified:

ΔT (temperature) $\Leftrightarrow \Delta V$ (potential difference)
 R_t (thermal resistance) $\Leftrightarrow R_e$ (electrical resistance)
 P (heat in Watts) $\Leftrightarrow I$ (current)

Likewise, the thermal counterpart for capacitance is the thermal capacitance. The thermal capacitance of the ground is defined as the amount of heat that can be absorbed per unit volume of the soil.

The product of the thermal resistance and the thermal capacitance is the thermal time constant.

One of the problems concerning underground cables, is the matching of the load cycle with the thermal time constant of the surrounding soil. Due to the thermal time constant it is possible to exceed the rated current of the cable for a short period of time during emergency and cyclic loading conditions.

The international standard for calculating the emergency cyclic current rating of cables is contained in IEC Document nr. 853 Parts 1 and 2. The methods in these standards apply only to cables in uniform soil (methods to deal with regions having different thermal characteristics are under consideration) and no account has been taken to represent the possible affects of moisture migration on the thermal resistivity of the surrounding soil. Furthermore it requires that the load cycle be substantially the same over a 24 h period. It also assumes that the thermal properties of the soil remains the same in time and space.

The ultimate aim of this study is to verify the existing models with respect to the South African (Highveld) climatic conditions. These models will then be modified to include effects such as layered soil and moisture migration.

The approach is to perform experiments on cable lengths under controlled conditions and then to investigate possible mathematical models to represent the measured data as accurately as possible.

2. EXPERIMENTAL WORK

To achieve above mentioned objective, an experimental set-up was designed consisting of a 1,5 x 1,5 x 3 m wooden box with the test cable running through it in the axial direction. Provision is made to fill the box with different types of soil.

As is shown in Figure 1, current is injected in the cable using three reversely connected current transformers. The cable forms the shorted secondary winding of the current transformers. The primary windings are connected in series so that the voltage is shared equally between them while the total current in the test cable can be increased to the desired level without saturating the current transformers.

A computer logging system, shown in Figure 2, was designed, built and tested to measure the temperature distribution in the soil surrounding the cable. The system makes provision for forty eight temperature measurements. The sensors will be placed at various points in the soil and the measurements will be used to verify the accuracy of the existing models.

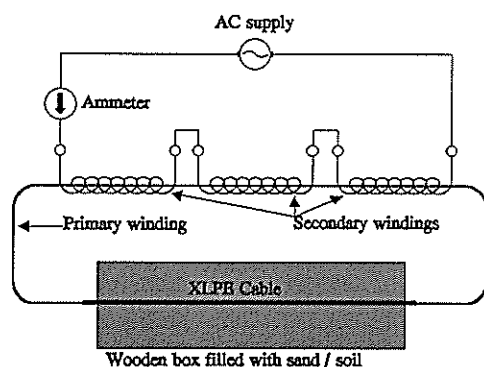


Figure 1: Cable buried in soil at a depth of 0,75m.

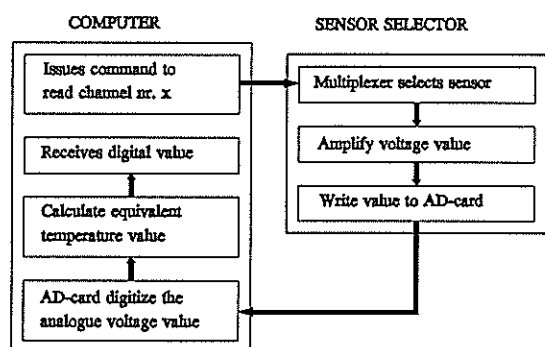


Figure 2: Block diagram of temperature logging system.

3. PRELIMINARY TEST RESULTS

At this stage only the experimental results for the cable in free air are available.

Figure 3 contains a graph that illustrates the thermal behaviour of the cable in air.

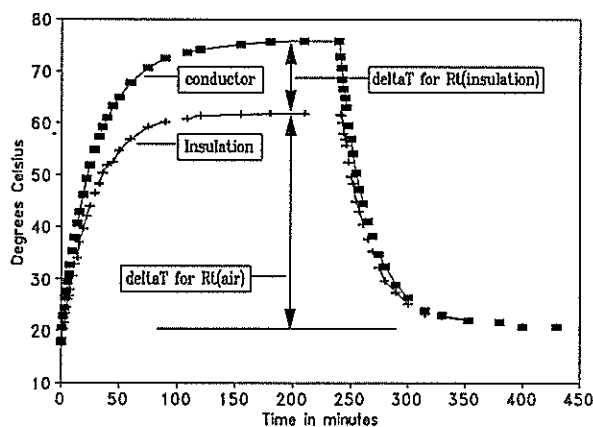


Figure 3: Thermal cycle of a cable in free air.

As can be seen from the graph, the temperature reached a maximum after about 4 hours. The power supply to the cable was terminated at this stage and the cooling of the cable was monitored for a further 3 hours and 20 minutes.

The two curves represent the temperature of the conductor and the temperature on the surface of the cable, respectively.

In Figure 4 a graph is shown, giving the maximum cable temperature as a function of the current flowing in the cable.

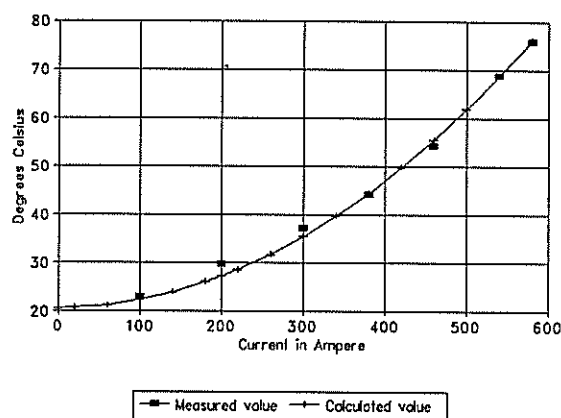


Figure 4: Maximum cable temperature as function of the current.

4. INTERPRETATION OF RESULTS

The following table contains a summary of the results for the experiment conducted with the XLPE cable suspended in free air, together with theoretical calculated values as detailed in Appendix A.

	Theory	Experiment
Copper temperature ($^{\circ}\text{C}$)	77,92	75,8
Cable temperature ($^{\circ}\text{C}$)	59,77	62,8
Temperature difference	18,15	13,0
$R_t(\text{air})$ (K.m/W)	0,809	0,864
$R_t(\text{insulation})$ (K.m/W)	0,378	0,268
Power dissipated (W)	47,935	48,34

As can be seen from the tabulated values, there is a very good correlation between the experimental and the theoretical results. The difference in the values obtained for the thermal resistance of the insulation, can be explained as follows:

The surface temperature of the cable was measured with a thermocouple that was held in position with a piece of masking tape. It is suspected that the tape

limited normal convection and radiation. Because of this piece of tape, the area in the immediate vicinity of the thermocouple had a higher temperature than the rest of the cable surface directly exposed to unhindered natural convection and radiation.

If it is assumed that the masking tape raised the temperature in area by 3.0°C , the modified value could then be calculated as follows:

$$\Delta T = 75,8 - 59,8 = 16,0^{\circ}\text{C}$$

$$R_{\text{(insulation)}} = \frac{16,0}{48,34} = 0,33 \text{ K.m/W}$$

These calculations show that an increase of about 5% in surface temperature lead to an 18% decrease in the value of $R_{\text{(insulation)}}$.

5. CONCLUSION

An experimental set-up was designed to analyze the thermal performance of the XLPE cables and to verify existing models.

Preliminary experiments, concerning cables in air, gave encouraging results. These results agree reasonably well with the theory.

Further work will involve tests with cables in soil and the development of theoretical models.

6. APPENDIX A

6.1 Calculations based on existing theory

Electrical resistance of copper (per unit length):

$$A = \pi \left(\frac{\text{diam}}{2} \right)^2 = \pi \left(\frac{14}{2} \right)^2 = 153,9 \text{ mm}^2$$

$$R_{DC}(20^{\circ}\text{C}) = \frac{kl}{A} = \frac{1,1,72\text{E}-8}{1,539\text{E}-4} = 0,1117 \text{ m}\Omega/\text{m}$$

The AC resistance of the cable at 90°C is:

$$R_{AC}(90^{\circ}\text{C}) = 0,1117 \cdot 1,275 = 0,14249 \text{ m}\Omega/\text{m}$$

The total power dissipated in the cable:

$$P = I^2 \cdot R_e = 580^2 \cdot 0,1425 \text{ m}\Omega = 47,935 \text{ W/m}$$

The thermal resistance from the copper conductor to the surface of the cable can be calculated with the aid of Figure 5.

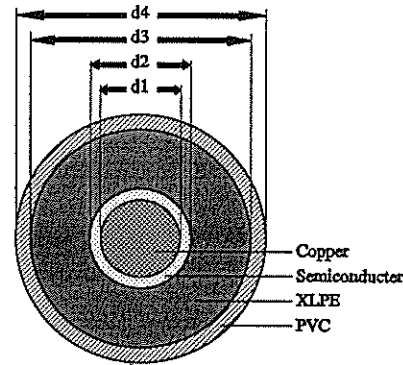


Figure 5: Cut through section of the 150 mm² cable.

The thermal resistivities for the various materials are obtainable from published work and are as follows [2]:

$$\rho_{\text{semicon}} = 3,5 \text{ K.m/W} \quad \rho_{\text{XLPE}} = 3,5 \text{ K.m/W}$$

$$\rho_{\text{PVC}} = 6,0 \text{ K.m/W}$$

The thermal resistance of the insulation is:

$$\begin{aligned} R_i &= \frac{\rho_{\text{semicon}}}{2\pi} \ln \frac{d_2}{d_1} + \frac{\rho_{\text{XLPE}}}{2\pi} \ln \frac{d_3}{d_2} + \frac{\rho_{\text{PVC}}}{2\pi} \ln \frac{d_4}{d_3} \\ &= \frac{3,5}{2\pi} \cdot \ln \frac{16,6}{14} + \frac{3,5}{2\pi} \cdot \ln \frac{24,1}{16,6} + \frac{6,0}{2\pi} \cdot \ln \frac{26}{24} \\ &= 0,378 \text{ K.m/W} \end{aligned}$$

The temperature drop over the insulation is:

$$\Delta T = P \cdot R_i = 47,935 \cdot 0,378 = 18,12^{\circ}\text{C}$$

Calculation of the thermal resistance for the ambient, can be done by using three resistances in parallel, one each for convection, radiation and conduction. It is known that air is a very poor conductor of heat and since the three resistances are in parallel, conduction can be ignored. The calculation for the remaining resistance, can be done as follows [2]:

$$R_{\text{(air)}} = \frac{1}{\pi d (f_c \alpha_c + f_r \alpha_r)}$$

α_c = transfer coefficient for convection W/K.m^2

α_r = transfer coefficient for radiation W/K.m^2

f_r, f_s = constants depicting the cable position

Calculation of the coefficients gives the following results:

$$\alpha_r = 7,145 \quad \alpha_c = 7,919 \quad f_r = 1 \quad f_c = 1$$

The outer diameter of the cable = 26,1 mm

$$R_{\text{f(air)}} = \frac{1}{26,1E-03 \cdot \pi (7,919,1 + 7,145,1)}$$

$$= 0,809 \text{ K.m/W}$$

With the thermal resistance of the air, the temperature on the outside of the cable is:

$$\Delta T = P \cdot R_{\text{f}} = 47,935 \cdot 0,809 = 38,779 \text{ K} \approx 38,7^\circ\text{C}$$

$$\begin{aligned} \text{Surface temp.} &= \text{Ambient} + \text{Temp. drop} \\ &= 21 + 38,779 \\ &= 59,779^\circ\text{C} \end{aligned}$$

Thus giving a copper temperature of:

$$\text{Temp. conductor} = 18,15 + 59,779 = 77,92^\circ\text{C}$$

6.2 Calculations based on experimental results

According to data sheets, the AC electrical resistance at 90°C is 0,1581 mΩ/m. Measurements revealed that from the cable the area of the copper is bigger by about 10%. The resistance then becomes

$$\begin{aligned} R_c &= \frac{0,1581}{(150 \cdot 110\%)} \cdot 150 \\ &= 0,1437 \text{ mΩ/m} \end{aligned}$$

The heat generated in the cable, amounts to

$$P = I^2 \cdot R_c = 580^2 \cdot 0,1437 = 48,34 \text{ W/m}$$

The copper temperature at steady state was 75,8°C and the temperature on the outside of the cable was 62,8°C.

This gives a value of

$$R_{\text{f}} = \frac{\Delta T}{P} = \frac{75,8 - 62,8}{48,34} = 0,268 \text{ K.m/W}$$

for the thermal resistance of the insulation.

From these figures the thermal resistance of the air can be calculated as

$$R_{\text{f(air)}} = \frac{\Delta T}{P} = \frac{62,8 - 21,0}{48,34} = 0,864 \text{ K.m/W}$$

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A Comparison between Simulated and Experimental Results for a Field Oriented Controlled Induction Machine

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Abstract

Field Oriented Control (FOC) is recognised as the premier method of AC machine control, in that it enables an AC machine to achieve a dynamic performance equivalent to that of a separately excited DC machine. Since an analogue implementation of the FOC algorithm is complex and costly a digital controller implementation is usually used.

This paper discusses the simulation and implementation of a digitally controlled FOC. The complete system is simulated in software and then practically implemented using a high performance digital controller in the laboratory. Results captured from the practical system are presented and compared with those of the simulated FOC system as a conclusion to the paper.

Introduction

The attributes of the induction motor such as ruggedness, reliability, low cost and high power to weight ratio have led to the increased use of induction machines in industry in recent years. Consequently, the complex dynamics of the induction machine and the algorithms necessary to achieve speed and position control of the machine have been the subject of considerable research. Based on this research FOC is now the universally accepted method for high performance induction machine control. In particular, the indirect method of FOC is considered to be the most practical scheme since it does not require the measurement of magnetic fluxes within the induction machine, and consequently standard machines may be used.

This paper discusses the simulation and implementation of a FOC system and compares the results captured from a practical system with those from the simulated system.

FOC Simulation

Field Oriented Control (FOC) is a technique whereby the complex non-linear and cross-coupled structure of an induction machine can be controlled to behave in a linear fashion, essentially very similar to that of a

DC machine. The principle of indirect FOC is well known [1], and only a brief description is given here.

Using the d-q model of the induction machine the torque is given by[1]

$$T_{em} = \frac{L_m}{L_{22}} (\lambda_{dr} i_{qs} - \lambda_{qr} i_{ds})$$

The aim of FOC is to control the torque, produced by the induction machine, by establishing and maintaining an explicit angular relationship between the stator current vector ($\underline{i}_s$) and the rotor flux vector ($\underline{\lambda}_r$) as shown in Fig. 1. This explicit angular relationship may be achieved by regulating the slip of the induction machine to that value which causes the rotor flux vector $\underline{\lambda}_r$ to become aligned with the orthogonal d-axis component (i_{ds}) of the stator current vector ($\underline{i}_s$). This value of slip is given by[1]

$$s\omega_e = \frac{L_m R_2}{L_{22} \lambda_{dr}} i_{qs}$$

Under these conditions of field orientation the q-axis component of the rotor flux (λ_{qr}) is zero and the torque of the induction machine IM simplifies to

$$T_{em} = \frac{L_m}{L_{22}} \lambda_{dr} i_{qs}$$

The induction machine torque is now a direct function of the q-axis stator current i_{qs} and the d-axis rotor flux λ_{dr} . Thus if i_{ds} can be held constant, the torque (and subsequently the speed) can be controlled by regulating the q-axis stator current i_{qs} .

Thus to control the induction machine it is necessary to regulate i_{ds} and i_{qs} . However these currents are not directly accessible. Hence external current controllers have to be designed that manipulate the stator voltages (v_{ds} and v_{qs}) in such a way that i_{ds} and i_{qs} follow the reference currents (i_{ds}^* and i_{qs}^*) as closely as possible. To achieve this the FOC control system shown in Fig.

2 has been designed. Fig. 2 shows the required q-axis and d-axis current controllers, as well as the FOC controller and precompensating controller. To enable speed and position control of the induction machine Fig. 2 also includes a speed control loop and a position control loop. Further details of the field oriented control system design can be found in [2].

Software programs written using CASE2 [3] were developed to simulate the system shown in Fig. 2. CASE2 enables quick development of the system model, however the simulations are time consuming, and consequently the system simulation was rewritten using C++ to speed up the simulation process. These simulations were performed using the parameters of an actual practical system (these parameters are shown in Appendix A). The results captured from these simulations will be discussed in a later section when they are compared with the results captured from the practical system described in the following section.

The Practical System

The practical FOC system was implemented using the hardware shown in Fig. 3. The main components shown in Figure 3 are as follows:

Caged Rotor Induction Machine: A 4 pole pair caged rotor induction machine was chosen. The parameters of this induction machine are given in Appendix A.

Embedded Programmable High Speed Controller(EPHSC): The EPHSC contains a T800 transputer as well as various I/O interfaces for the IM environment (eg. A/D, PWM, and Shaft Encoder interfaces), and is capable of providing the necessary processing requirements for the computation of the FOC scheme. A detailed description of the EPHSC can be found in reference [4].

Inverter: The 3 phase inverter consists of six IGBT power devices and their associated drive / isolation circuits. The PWM switching signals for the inverter are generated by the PWM I/O interface of the EPHSC, this interface contains the Hanning PWM ASIC [4].

Speed Sensor: An incremental type shaft encoder was used to sense the position of the IM shaft. The EPHSC provides an interface to determine the speed of the machine shaft from this encoder.

Current sensors: Two hall effect current sensors were connected between the induction machine

and the inverter. The current sensors produce a voltage which is proportional to the current flowing and this voltage was sampled by the A/D interface of the EPHSC for use by the FOC algorithm.

DC Voltage Source: The DC voltage source (275V) was supplied by a battery system in the laboratory.

The Practical System Software: The EPHSC was programmed using the Occam parallel processing language. The Occam language is based on the concepts of Communicating Sequential Processes (CSP) and enabled the FOC algorithm to be described as a number of communicating sequential tasks or processes.

Comparison Of the Practical System Result with the Simulated System Results

In this section the results captured from the practical FOC system are shown and these results are compared with those of the simulated system.

1. Practical and Simulated system response to a step change in the speed reference.

A step input was applied to the speed reference of the practical FOC system and the results captured are shown in Figs. 4 - 8 along with the corresponding results from the simulated system.

The speed controller gains in the practical system were set to the values shown below (these are the same values as used for the simulated system and were derived from theoretical calculations based on the motor parameters given in Appendix A).

$$K_{ps} = 7$$

$$K_{is} = 15$$

Prior to the application of the step change to the speed reference, the IM was prefluxed (λ_{dr}) by applying $i_{ds} = 4.9(A)$ for approximately 0.2(s). A step change in speed from 0-314(rad/s) was then applied to the speed reference input. Fig. 4 shows that the speed response of both the simulated and the practical system rises linearly to 314(rad/s) with no oscillations and negligible overshoot. The q-axis current (which is representative of the torque in the machine see Fig. 8) developed during this speed transient is shown in Fig. 5 for both the simulated and practical systems. Again the responses are similar except for a noise component present on the signal captured from the practical system. Fig. 5 shows that as the FOC system responds to the speed transient the q-axis current rises rapidly to the limited value of 30(A) and remains there until the speed of

the machine reaches the commanded value of 314(rad/s), at which point the q-axis current rapidly falls to the no-load level. The d-axis current flowing in the induction machine is shown in Fig. 6 and approximately constant at the required value of 4.9(A), however both the simulated and practical d-axis current responses display considerable noise. The occurrence of this noise stems from assumptions made during the design of the d-axis current controller- a detailed explanation of which can be found in [2]. Fig. 7 shows the q-axis component of the rotor flux λ_{qr} for the simulated system only (the measurement of this quantity in the practical system is extremely difficult). It can be seen that the value of λ_{qr} remains approximately equal to zero and consequently that field orientation has been achieved within the simulated system. Furthermore, the similarities between the responses of the simulated and practical system in general suggest that field orientation is also occurring within the practical system (unfortunately this difficult to verify by measurement).

2. Practical and Simulated system response to step change in the position reference.

The position controller gain was selected to be

$$K_{pp} = 14$$

for both the simulated and the practical systems. To evaluate the response of the position control loop a step change was applied to the position reference input and the responses shown in Fig. 9-12 were captured from both the simulated and the practical system. Fig. 9 shows the step change in the position of the induction machine shaft from 0-314(rad/s), whilst Fig. 10 and Fig. 11 respectively show the d-axis current (i_{ds}) and the q-axis current (i_{qs}) components present in the machine during the position transient. The simulation results compare favourably with the responses captured from the practical system and it can be clearly seen that the d-axis current (i_{ds}) remains virtually constant whilst the q-axis current component (i_{qs}) changes in response to the demand for acceleration and braking torque (braking torque is indicated by the negative values of i_{qs} in Figure 11). Figure 12 shows that the q-axis flux $\lambda_{qr} = 0$, the system is controlled under FOC.

Conclusion

This paper has briefly discussed the simulation and implementation of a field oriented controlled induction machine. Results were presented and compared from both the simulated system and the practical system and it was found that the simulated system response accurately matched that of the practical system.

Appendix A: Induction Machine Parameters.

Power	$P = 3 \text{ (KW)}$
Voltage	$V = 220 \text{ (V)}$
Current	$I = 14 \text{ (A)}$
Inertia	$J_e = 0.46 \text{ (kg.m}^2\text{)}$
Stator resistance	$R1 = 0.3432 \text{ (}\Omega\text{)}$
Rotor resistance	$R2 = 0.6564 \text{ (}\Omega\text{)}$
Stator leakage reactance	$X1 = 0.9811 \text{ (}\Omega\text{)}$
Rotor leakage reactance	$X2 = 1 \text{ (}\Omega\text{)}$
Mutual reactance	$Xm = 32.4277 \text{ (}\Omega\text{)}$

Reference:

- [1] F Blaschke, "The Principle of Field Orientation as applied to the new TRASVECTOR Closed-Loop Control System for Rotating Field Machines", Siemens Review, Vol.34, May 1972, pp. 217-220.
- [2] Hao B, "Implementation Of Field Oriented Control for a VSI fed Caged Rotor Induction Machine", MSc Thesis to be submitted, University of Natal, February 1994.
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- [4] Woodward D R, "Transputer based Embedded Programmable High Speed Controller for AC drives", MSc Thesis, University of Natal, November 1992.

Figure:

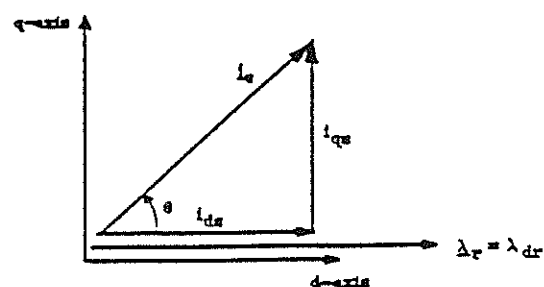


Fig. 1 Conditions for correct field orientation

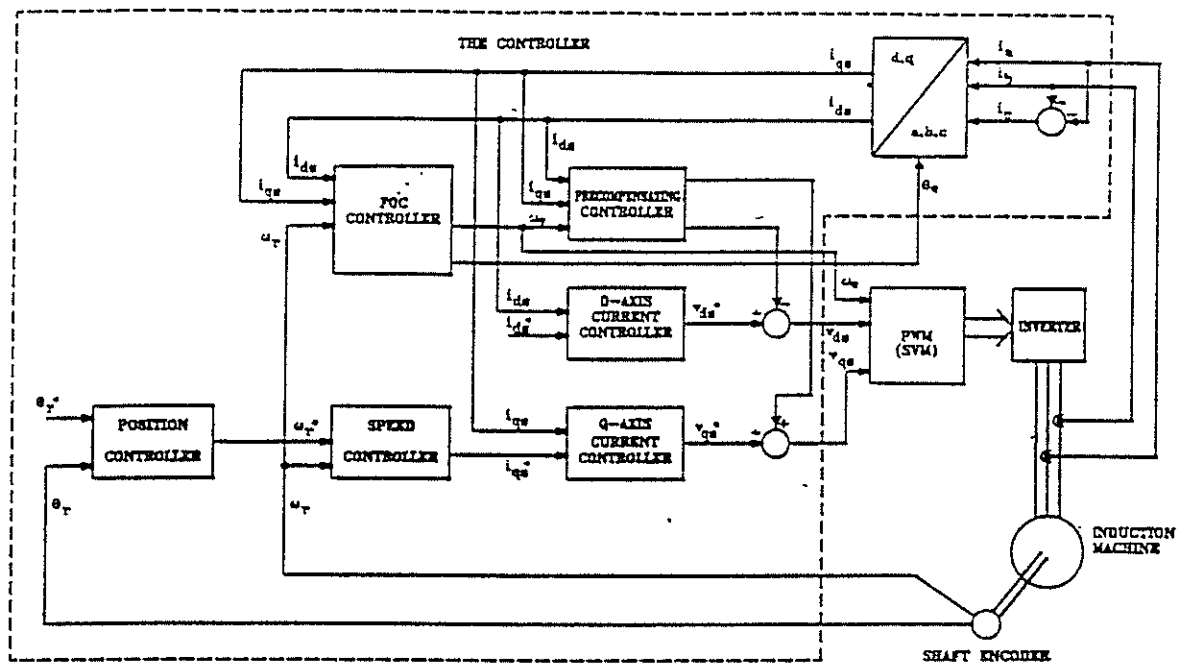


Fig. 2 Typical drive structure of a motion control system using FOC

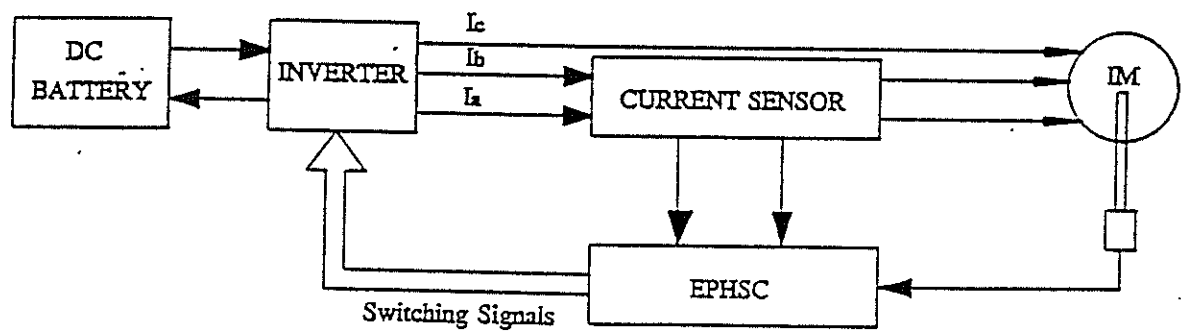


Fig. 3 The implement system of FOC

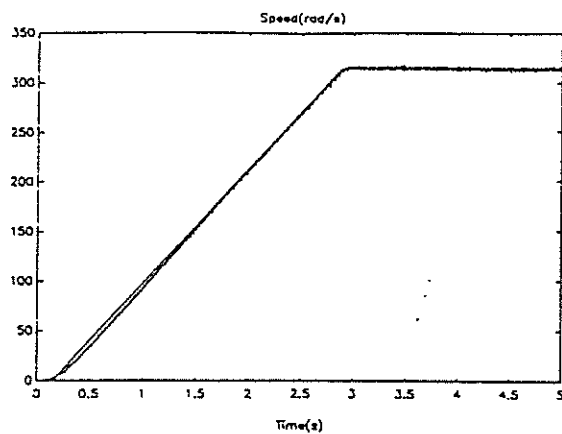


Fig. 4 Speed(rad)

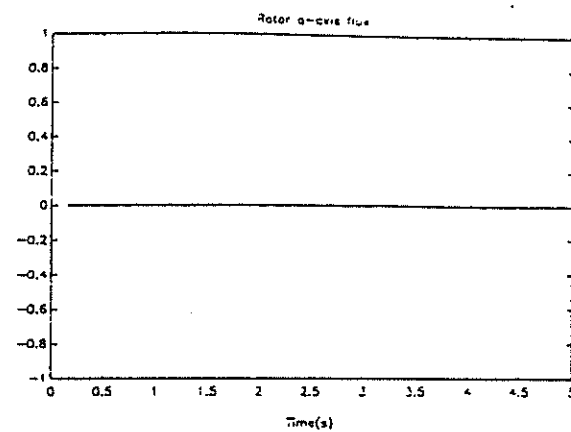


Fig. 7 Rotor q-axis flux(Wb)

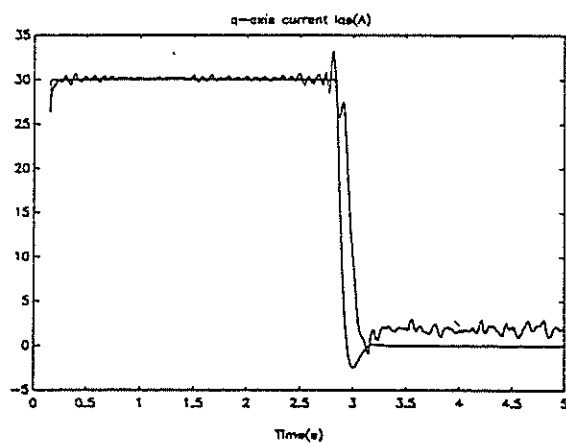
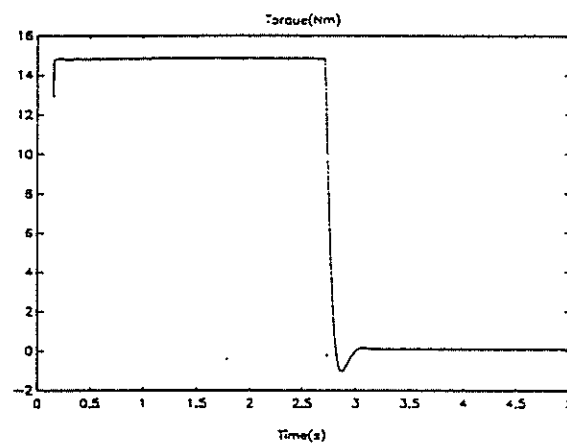
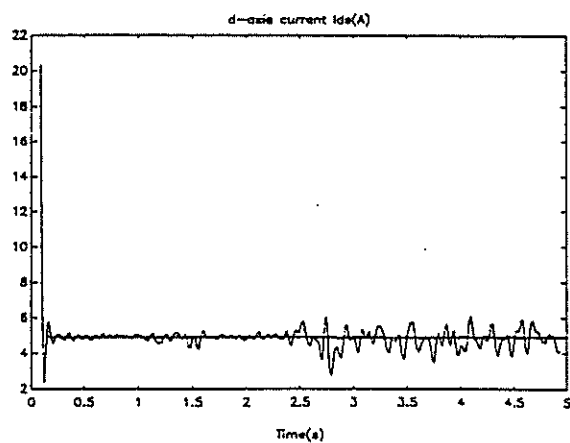
Fig. 5 Q-axis current I_q (A)

Fig. 8 Torque(Nm)

Fig. 6 D-axis current I_d (A)

Response of the speed ctrl system with 30(A) current limit

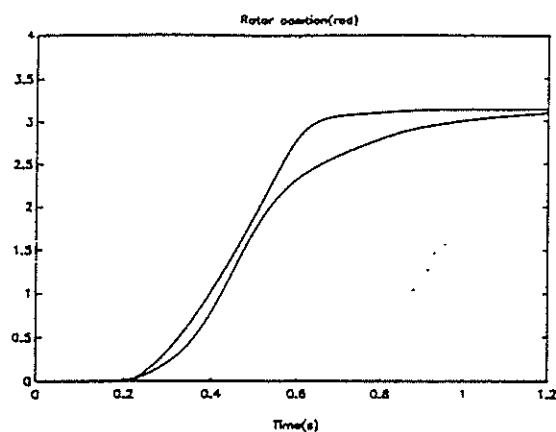


Fig. 9 Position(rad)

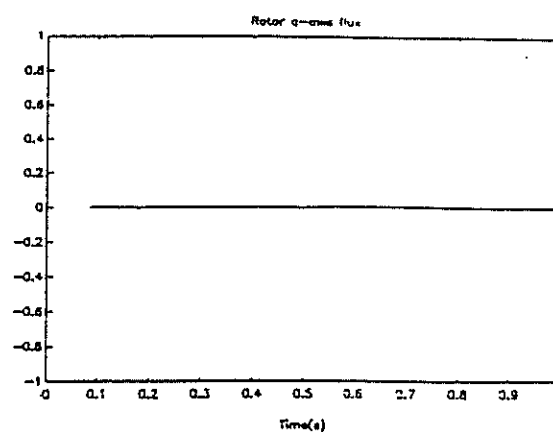
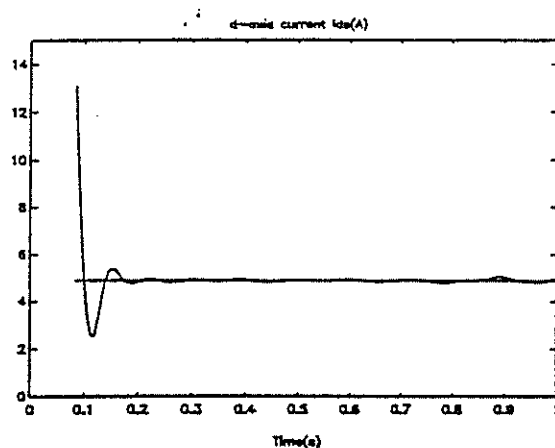
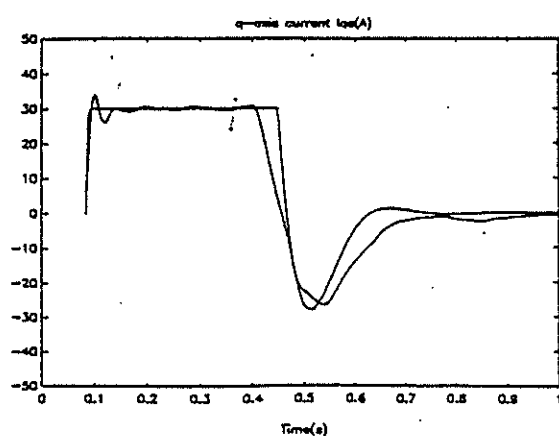


Fig. 12 Rotor q-axis flux(Wb)

Fig. 10 D-axis current I_d (A)Fig. 11 Q-axis current I_q (A)

Response of the position control system of the simulation and the practical (0-3.14(rad))

DYNAMIC PROPERTIES OF THE SWITCHED RELUCTANCE MOTOR

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1. Abstract

In recent years, a multitude of articles concerning the Switched reluctance motor have been published. Yet not one of these provide the newcomer to the field, with a concise guide to the effect of the motor parameters on the dynamic properties of the motor. The purpose of this paper is to fill this gap in the literature. The dynamic properties considered here are: Torque-pulsations, average torque and operating speed.

2. Introduction

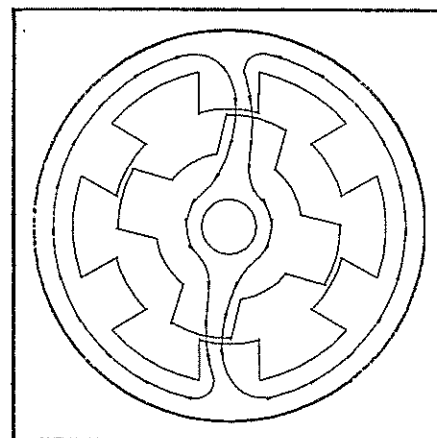
The Switched-reluctance motor (SRM), as shown in figure 1, is a doubly-salient structure with an uneven number of stator and rotor poles. The motor works on the principle that any magnetic circuit will always revert to a position of minimum reluctance. The reluctance of the magnetic path for the rotor position shown in figure 1 a. is higher than that for the rotor position shown in figure 1b. If the correct stator coils are excited the rotor will move from position 1 (shown in figure 1 a.) to position 2 (shown in figure 1b.). By switching the coils wound around each stator pole in the correct sequence, (determined by the position of the rotor), rotation and positive torque can be obtained. References [1] and [2] offers a more comprehensive treatment of the operating principles and in the rest of this paper it will be assumed that the reader is familiar with it.

3. Dynamic properties

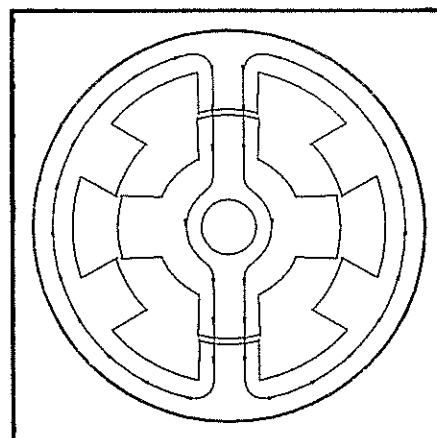
Various design procedures for the Switched-reluctance motor have been published [3,4]. But before one gets involved in the rather complicated theory proposed in the design procedures, it is advisable to familiarize oneself with the effect of the various motor parameters on the dynamic operation of the motor.

3.1. Torque

One of the most important specifications in the design of a Switched-reluctance motor, (and indeed any other motor), is the torque required. Not only the magnitude but also various other requirements such as the amount of torque pulsations that can be tolerated. The most basic of these requirements is that of constant torque production.



a.)



b.)

Figure 1: Switched-reluctance motor configuration.

Constant torque production in the Switched-reluctance motor, can only be achieved if at least one pair of stator and rotor poles are overlapping, and thus capable of producing torque, at any rotor position. This will naturally place some sort of a constraint on the stator and rotor pole arcs. These geometric constraints has been neatly summarized by Lawrenson [2] and are shown below.

$$\text{LCM} (N_S, N_R) = qN_R \quad (1)$$

$$\min (\beta_S, \beta_R) > 2\pi/qN_R \quad (2)$$

$$\beta_S \leq 2\pi/N_R - \beta_S \quad (3)$$

Where: LCM = Least common multiple

min = minimum

N_S = Number of stator poles

N_R = Number of rotor poles

q = Number of phases

β_S = Stator pole arc

β_R = Rotor pole arc

A multitude of formulas for the torque produced by the Switched-reluctance motor has been proposed [1,3,5]. The most general formula [1] is:

$$T = 1/2 i^2 dL/d\theta \quad (4)$$

Where: T = Instantaneous torque

i = Instantaneous current in the stator winding

L = Inductance

θ = Angle of rotation

From equation 4 it can be seen that the torque produced by one set of stator and rotor poles are dependent on the current that flows in the stator windings. Since the stator windings have a certain inductance, the current takes a finite time to increase from zero to maximum, when the windings are connected to the supply. Because of this inductance the current can also not decrease immediately to zero when switched off. This results in a torque wave form of a similar shape as is shown in figure 2. It must be noted here that the wave form in figure 2 only serves to illustrate the general shape of the torque pulse and is not intended to represent any motor configuration in particular. The stator windings of one phase of the motor are

usually connected to the supply as the stator and rotor poles (for this phase) starts to overlap. The supply is disconnected at the moment that the two poles are in alignment. It is important that the supply must be disconnected and the current must have decreased to an acceptably low level before the poles start to move out off alignment. If this is not adhered to a negative torque will be produced that would lower the average positive torque produced by the motor as a whole.

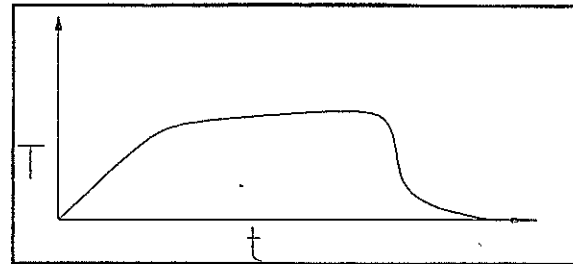


Figure 2: Torque wave form.

Each phase of the motor produces a torque pulse similar to the one shown in figure 2. As the motor rotates, different sets of stator and rotor poles will come into alignment at different times*. With proper switching of the stator windings, each phase will produce a torque pulse similar to the one shown in figure 2. These pulses will be displaced in time and will combine to produce the average torque of the motor. Figure 3 shows this torque pulse-train as well as the instantaneous torque curve that it produces.

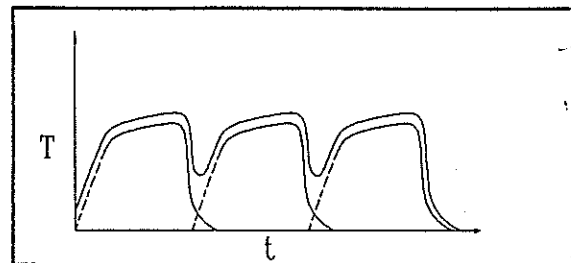


Figure 3: Torque pulse-train and instantaneous torque.

From figure 3 it can be seen that the torque pulsations, or torque ripple, is dependent on the amount of overlap between the individual torque pulses. A design based on equations 1,2 and 3 will ensure that there is always some

* This is a result of the uneven number of stator and rotor poles[1,2].

amount of overlap. Looking at figure 3, it might seem that the problem of torque pulsations, inherent to this type of motor, can easily be solved by designing a motor with the correct amount of overlap of the torque pulses. According to equations 1.2 and 3, this would imply bigger stator and rotor pole arcs. But the torque produced is dependent on the rate of change of the inductance of the magnetic path, $dL/d\theta$ (eq. 4). Figure 4 shows the inductance profile as the rotor rotates around the air gap.

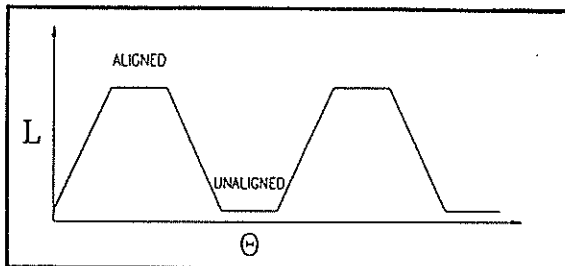


Figure 4: Inductance profile.

From figure 4 it can be seen that $dL/d\theta$ is dependent on the difference between the maximum and minimum values of the inductance*. But:

$$L = N^2/\mathfrak{R} \quad (5)$$

Where: L = Inductance

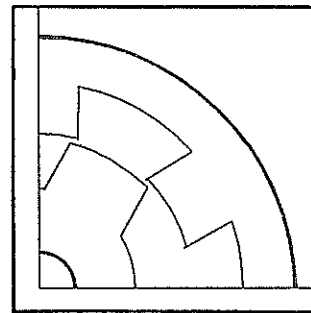
N = Number of turns

$\mathfrak{R}$ = Reluctance of the magnetic path

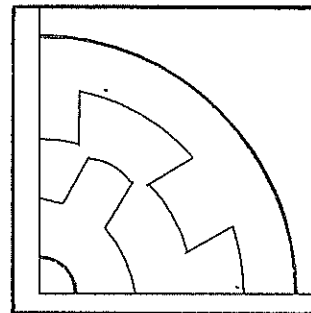
Figure 5 a. shows a motor with large values of β_s and β_r and figure 5 b. a motor with smaller values of β_s and β_r . The magnetic path of the configuration shown in figure 5a., clearly has a lower reluctance, (because of the bigger overlap), than the configuration in figure 5b. According to equation 5 this will result in a higher value of minimum inductance for the motor with large values of β compared to the motor with lower values of β . The inductance for the aligned pole position of both motors are roughly the same. We can therefore deduce that the motor of figure 5a. will have a lower value of $dL/d\theta$ than the motor of figure 5b.

According to equation 4, a lower $dL/d\theta$ will result in a lower torque. From this we can conclude that: Although we can reduce the

torque pulsations by enlarging β_s and β_r , we will at the same time be reducing the maximum value of the torque produced.



a.) Large values of β



b.) Small values of β

Figure 5: Comparison of motor configurations (unaligned poles).

It is obvious that a tradeoff has to be made between the amount of torque pulsations that is acceptable and the magnitude of the torque required. Other methods to reduce the torque pulsations by means of the power electronic converter has been proposed but these fall outside the scope of this paper.

3.2. Operating speed

Another important specification at the design of any motor is the operating speed. The speed of the Switched-reluctance motor can be controlled by limiting the magnitude of the phase current. Since the torque of the motor is dependent on the current (eq. 1), limiting the current would imply a limiting of the torque. At relatively low speeds this procedure works well. At high speeds however, the switching frequency is increased, and thus the on-time of each phase is drastically reduced. The motor speed reaches a limit when the time available for the rise of current in the winding, limits the current to a value that will produce the torque

* If we compare two motors with the same stator diameter, rotor diameter, N_s and N_r .

required to overcome friction and the load torque. At this specific speed of rotation the motor will drive the load at a certain stable speed. At very high speeds it becomes impossible to inject any significant current into the windings. This is the explanation for the shape of the torque speed curve of this motor, as shown in figure 6 (torque decreases with an increase in speed).

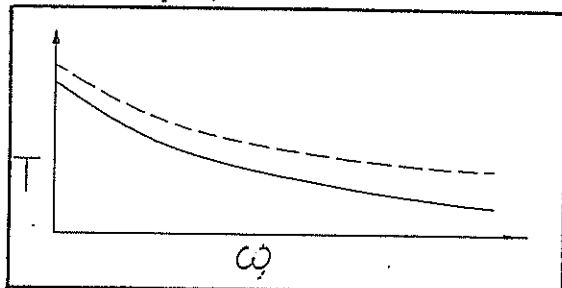


Figure 6: Torque speed curve.

If high speed operation is required, the designer must attempt to lower the inductance of the magnetic path. This can be achieved by lowering the number of windings on each stator pole (eq. 5). This will lower the inductance but it will also lower the flux (ϕ) set up in the motor since:

$$\phi \propto Ni \quad (6)$$

If we want to decrease N , we will have to increase i accordingly. This might lead to overheating problems if the high current is not designed for. A reduction in the inductance would, however, have the benefit of a higher operating speed at a certain operating point (see the dotted curve on figure 6), because of the higher rate of rise of the current in the windings. It must be mentioned here that eq. 6 only holds for low values of saturation. As soon as the iron starts to saturate a further increase in current would not cause a significant increase in the flux of the iron magnetic circuit. A reduction in the inductance is however still beneficial since it will cause a shorter rise time for the current during the initial stages after switch-on. A further method to reduce the inductance is to connect all the windings, on each stator pole of a phase, in parallel. This would invariably lead to the switching of very high currents in the converter. The author is currently working on different motor geometry's

to try and minimize inductance while maximizing torque.

4. Test results

A Switched reluctance motor utilizing a inductance motor stator was designed and tested by the author. The power derived from this configuration was much less than that achieved by other authors, using purposely designed stators [2,4]. The explanation for this is that the difference between the maximum and minimum inductances for the induction motor stator, is very low due to the partially closed slots in the induction motor stator. These results confirm the principles stated above. Further investigations utilizing a finite element method package also confirmed the stated design principles.

5. References

- [1] Miller, T.J.E. : "Brushless Permanent-Magnet and Reluctance Motor Drives" New York, Oxford University Press, Chapter 7, 1989.
- [2] Lawrenson, P.J. ; Stephenson, J.M. and Blenkinsop, P.T. : "Variable-speed switched reluctance motors" IEE Proceedings, vol. 127, Pt. B, pp. 253-265, 1980.
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- [4] Miller, T.J.E. and McGlip, M. : "Nonlinear theory of the switched reluctance motor for rapid computer-aided design" IEE Proceedings, vol. 137, Pt. B, no. 6, pp. 337-347, November 1990.
- [5] Miller, T.J.E. : "Converter Volt-Ampere Requirements of the Switched Reluctance Motor Drive" IEEE Transactions on Industry Applications, vol. 1A-21, no. 5, pp. 1136-1144, Sept./Oct. 1985.

THE MEASUREMENT OF THE CURRENT TRANSITIONS IN THE ARMATURE CONDUCTORS UNDERGOING COMMUTATION IN A DIRECT CURRENT MACHINE

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A.S.Meyer C.F.Landy
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ABSTRACT

A universal laboratory electrical machine was connected to function as a D.C. machine. Electrical access to the normally closed armature windings of this machine could be obtained via slip-rings. It was therefore possible to extend the armature conductors between two adjacent commutator segments to a point where a Hall effect current probe could be employed to measure the current in these conductors while the machine was under operating conditions.

The current transitions thus measured and observed were used to assess the commutation performance of the machine as a function of interpole action which was brought about by exciting certain auxiliary windings on the machine.

The conditions for over, under, and perfect commutation were measured and observed in this way and compared to the standard textbook postulations of the commutation process.

Additionally the role that brush and brush contact resistance plays in the commutation process was clearly observable from these tests.

An unexpected trend of the current in the conductors undergoing commutation in the absence of interpole action has prompted a new theoretical investigation into the commutation process.

INTRODUCTION

All explanatory texts, too numerous to mention, dealing with the commutation process put forward the standard notion that during this process the current in the coil undergoing commutation should ideally undergo a complete reversal during the commutation period. Deviations from this ideal are those conditions whereby a complete reversal of current does not occur during the commutation period, or when more than a complete reversal occurs. These conditions are referred to as under- or over commutation respectively.

These under, over, and ideal commutation conditions are inferred from the well known pilot brush

potential curves, an example of which is shown in figure 1. However, pilot brush potential curves provide no information whatsoever as to the form of the actual current transition in the armature conductors undergoing commutation. Diagrams depicting these current transitions that appear in the various texts are therefore based on speculation. Equally speculative are terms such as straight-line commutation and sine-wave commutation, sometimes used in these texts to describe the form of these current transitions. This paper describes how the actual current transitions in the armature conductors undergoing commutation were actually measured and the results of such measurements.

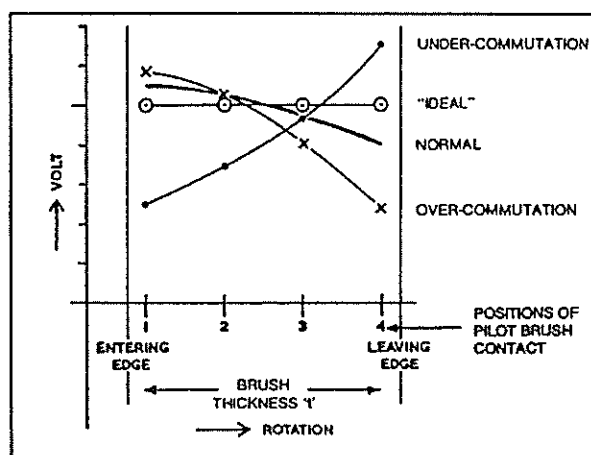


Figure 1: Pilot brush potential plots showing how commutation performance is inferred from the plots.

GAINING ACCESS TO ARMATURE WINDINGS OF A D.C. MACHINE

The armature windings of a D.C. machine form a closed circuit. However, a certain type of machine, specifically constructed for laboratory experimentation, was found to have facilities whereby the normally closed armature circuit could be opened to permit the insertion of measuring instruments. Naturally, access to an armature circuit conductor at

any point implies that two commutator segments exist at either end of that conductor. If the armature is lap wound, as this one was, then these two armature segments are adjacent to one another meaning that one brush will take part in the commutation process. This would not have been so for a wave wound armature but, in either case, access to an armature conductor permits the current to be measured in that conductor. And, during the course of one revolution of the armature, the current transition of the commutation process can be observed and measured with the appropriate instrumentation and circuit connections.

THE MEASUREMENT PROCEDURE

The two ends of the open-circuited armature winding of the test machine were brought up via slip-rings to two stationary terminals. In order to restore the continuity of the armature these terminals were short-circuited with a short length of conductor. A Hall-effect current probe was connected around this short-circuiting conductor. The current probe had a frequency response in excess of 5 MHz making it possible to detect the expected current transitions with considerable accuracy. The output of the probe's conditioning amplifier was connected to a storage oscilloscope with facilities to dump the screen image of the current trace to a printer.

To minimise the effects of anomalous brush resistance the commutator was dressed with a brush bedding stone and the machine was run until as smooth a current trace as possible was observed on the oscilloscope.

Appropriate adjustable interpole action was obtained by independently exciting certain auxiliary windings provided on the test machine for this purpose.

RESULTS

Figure 2 shows an oscilloscope trace of a measured current transition that can reasonably be described as representing perfect commutation. This is characterised by the absence of instantaneous changes in current and the consequential attempted instantaneous change in flux linkages.

Note that it appears from figure 2 that the commutation process is essentially, what textbooks would call, straight-line commutation during the major portion of the commutation period. However, at the beginning and end of the transition, the trace is rounded. This could possibly be due to the expected relatively small portion of the brush being in contact with the commutator segment and the

increased contact resistance arising there-from.

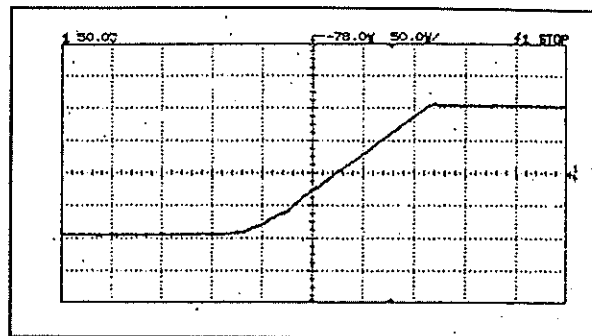


Figure 2: Perfect commutation characterised by the absence of instantaneous current transitions.

Increasing the current responsible for the interpole action had the expected effect of causing over commutation. This is obvious from the current trace shown in figure 3.

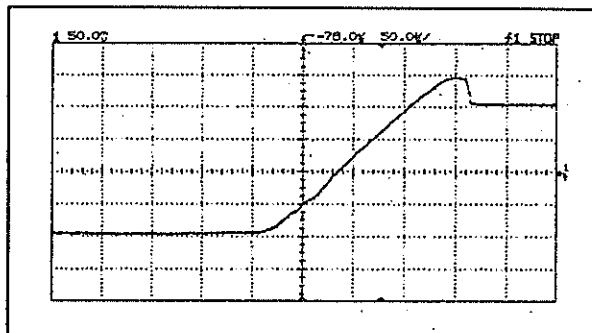


Figure 3: Over commutation

Figure 3 could nearly be regarded as a textbook illustration of over commutation except for the rounding of the current transitions, particularly at the end of the commutation period.

The instantaneous current transition at the very end of the commutation period is the undesirable consequence of over commutation. This instantaneous current transition can be made larger, by merely increasing the interpole current, to a point where commutator sparking would be likely to occur.

Decreasing the interpole current gave the current trace shown in figure 4. The instantaneous current transition at the very end of the commutation period

clearly illustrates the undesirable consequences of under commutation. The hump in the middle of the trace can only be due to some unquantifiable anomalous brush contact resistance. Otherwise this is, once again, a textbook illustration of under commutation.

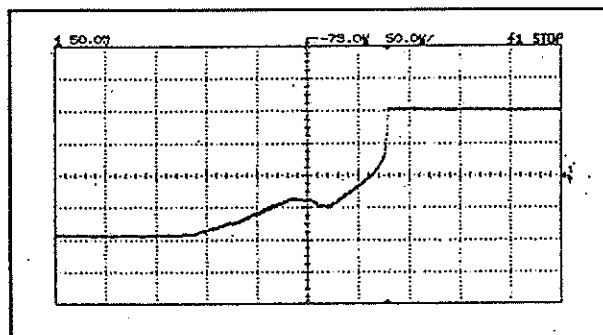


Figure 4: Under commutation

To obtain the trace shown in figure 4 the interpole current was only marginally reduced. This marginal reduction has caused an instantaneous current transition from about zero amps to the reverse full load current at the end of the commutation period. Taking this to the extreme, and decreasing the interpole current to zero yielded the unexpected current transition as shown in figure 5.

No explanatory text on the commutation process puts forward a commutation current transition as is illustrated in figure 5.

Moreover, this current transition cannot be inferred or postulated from pilot brush potential curves. The

excursion of the current transition beyond the load current at the onset of the commutation period suggests a hitherto unexplained effect, possibly due to armature reaction. This unusual behaviour of the current transition in the absence of interpole action could only have been revealed by the methods employed to observe the commutation process as detailed in this paper.

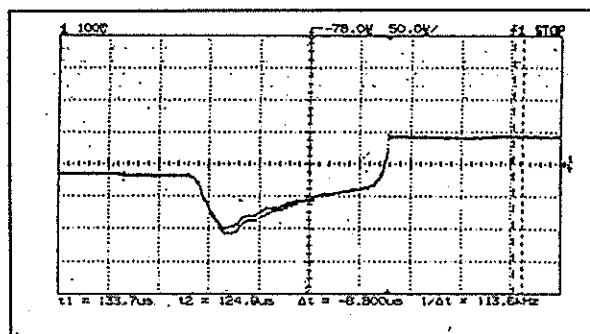


Figure 5: The current transition during the commutation process with zero interpole current.

CONCLUSIONS

By being able to measure and observe the actual current transitions occurring during the commutation period has revealed that our understanding of the commutation process is not complete. Explaining the reasons for the current transition shown in figure 5 has prompted the authors to undertake a considerable research effort with a view to developing a method of predicting these observed current transitions.

INTERACTION BETWEEN TWO LOOSELY COUPLED SYNCHRONOUS MOTORS FOR A DRAGLINE SYSTEM

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ABSTRACT

When General Electric moved its manufacturing capability of synchronous machines from Schenectady to Erie, the motor design changed. This study investigates whether the different motors may be run in parallel on a dragline. Initial investigations considering the operation of the motors running separately showed that potential problems may arise. This work simulated the transient operation of the motors in a complete system. It is shown that the system remains stable and the effect of the two designs is negligible on the dynamic response of the system. However, in order to have complete confidence in operating the two motors in parallel, controlled site tests should be performed.

1. INTRODUCTION

Draglines perform a critical role in clearing the overburden of open cast mines. It is therefore essential to have a high availability and short down times. Therefore spares of the essential motors are normally kept in the stores of the mines. Two of the important motors found on a dragline are the 3000 HP (2235kW) synchronous motors which are manufactured by General Electric.

Three mining houses share a common pool of spares. It was necessary to determine whether two synchronous motors with different designs could be used together, to minimise the number of spares held.

Recently, the manufacturing facility of these synchronous motors was moved from Schenectady to Erie, and in the move the design of the synchronous motor was altered. This involved changes to the damper windings, the field resistance and the V curve of the machine.

The alterations made to the field resistance and V curves of the machines affect their static operation, and thus will not be dealt with in this report.

The change in design of the damping bars of the synchronous motors changed the rating of the starting current from 500% to 700%. In addition, the damping ratio increased from 4.3 to 16.5. Problems were

foreseen if two motors with differing settling times were run simultaneously. When load changes, the torque on the synchronous motor oscillates until it settles down to a steady state value. Two motors oscillating with different time constants could cause additional stresses in the system.

Initial investigation showed that if the motors were run separately, the torque oscillations would be out of phase after three seconds, causing some concern. Subsequently, a more detailed system has been modelled and simulated, which accounts for the interaction between the two synchronous motors, via the DC generators and motors, and the mechanical load.

This report considers the dynamic behaviour of the overall dragline system, and simulation techniques are used to ascertain whether any problems arise. The simulation environment used to examine the transient behaviour is described. The description of the models used is then presented. The results of the simulation are shown and their impact discussed.

2. THE SIMULATION ENVIRONMENT - CASED

The simulation environment used for the studies was CASED (Computer Analysis and Simulation of Electric Drives). This environment has been developed over many years and has been specially tailored for the simulation of motors and drives.

CASED works on a modular basis and allows for the interconnection of electrical supplies, electric machines, mechanical models and special user definable controllers. The system which was modelled is discussed in detail in the next section.

Advanced numerical solvers for the non-linear ordinary differential equations (ODE) have been implemented in CASED. These solvers have an algorithm which allows it to continuously vary the step length over which the ODE is being solved. This feature ensures the numerical accuracy of the simulation results. Special solvers have also been implemented for stiff-systems, that is systems in which the time constants vary by several orders of magnitude.

3. THE SYSTEM

A dragline has two synchronous motors driving six DC generators each. The generators are connected in a Ward-Leonard configuration to DC motors which perform swing, drag and hoist operations. For simplicity, figure 1 represents only one of these operations. Thus, only two DC generators are shown driven by each synchronous motor to perform, for example, the hoist function.

The four DC motors of each set are connected to a common load, and can be said to all have the same speed. For modelling purposes the four DC motors can be replaced by an equivalent single DC motor, which has the same armature current of the original motors but four times the armature voltage of the original motors. The models of the DC machines and the synchronous machines ignore the effect of saturation in the flux paths.

Static regulators are used to control the excitation on the field of a rotary exciter driving the fields of the synchronous motors. These monitor stator current and voltage of the motor, and adjust the field excitation to maintain a desired power factor in the motor. (This simulation controlled the power factor to unity.) A slightly simplified controller was used, in that the stability feed back loop gain was set to zero. This was because the non-linear characteristics of the controller were not sufficiently specified. The effect of this simplification should be small.

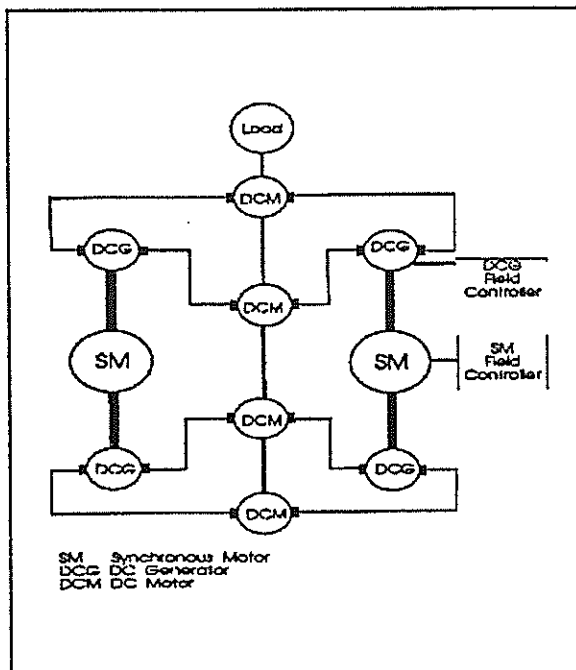


Figure 1 - Diagram of system modelled

Rated field current is produced in the DC motors. The field voltages of the generators are fed from a common exciter. The voltage output of the exciter is controlled so as to produce the required voltage and current in the DC generator. This controller ensures that both the armature voltage and the armature current are limited to within design values.

4. SIMULATION RUNS

Two types of simulations were performed: when the motors are running independently, and then the complete system as described in the previous section.

5. INDIVIDUAL MOTORS

Each of the two types of synchronous machine are run independently and loaded lightly with a load torque of 200 Nm (0.01 pu). The field controller is connected and the machine brought to a steady state condition.

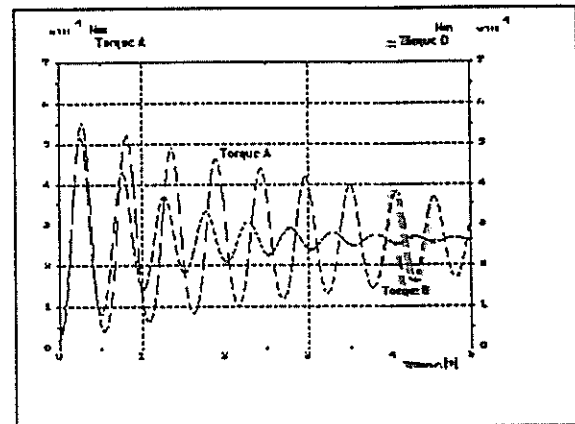


Figure 2 - Torque response to a step input for the two machines running independently

A sudden load torque of 26 kNm (1.2 pu) is then applied to the motors. By observing the airgap torque developed by the machine, the oscillations in the internal currents can be readily seen, as is shown in figure 2.

For the first half second, the torques produced by the two machines are almost identical. The peak torques produced by each machine are very similar, and therefore the peak stresses experienced by the mechanical system are comparable. From the figure, it can be clearly seen that the oscillations of the torque of the old design, Torque B, damp out much faster, as is expected. Further the natural frequency of the oscillations is higher than for the new design, Torque A. Therefore, at about 3 seconds the torques are anti-phase. This phenomenon may cause problems in the complete system, therefore it was decided to perform a complete system study.

6. SYSTEM STUDY

The complete system described earlier was modelled and simulated using CASED. Two studies were performed. In the first run, the DC motors were run up against a load torque such that the synchronous motors saw a reflected torque of 18 kNm (0.85 pu). In the second run, the DC motors experience a sudden increase in load torque, and the DC motors operate in current limit mode.

Run-up

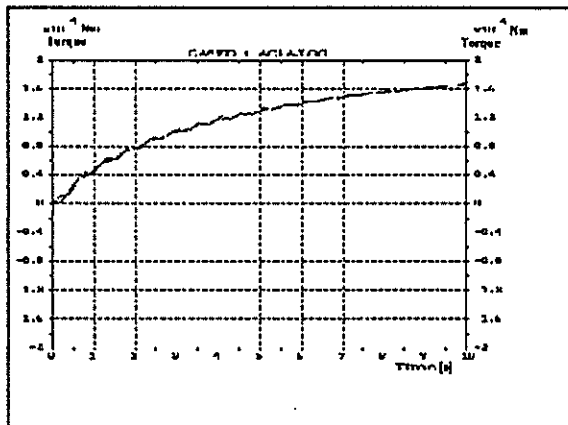


Figure 3 - torque on synchronous motors during DC motor runup

Figure 3 shows the torque traces of the two synchronous motors during the run-up. The two traces do differ from each other, but only by less than 850 Nm (0.04pu) and the oscillations die away, especially after 5 seconds. From this trace it is clear that no problems are anticipated during run-up.

Step load increase

The load torque on the DC motors is increased from 3.2 kNm to 6.5 kNm. The DC motors then go into a current limit, which was set at 5500 Amps, which is 20% above the rated stall current of the machine. This case, therefore, represents a worst case shock to the system. The trace of the developed airgap torques of the two synchronous motors are shown, in figure 4. Once again, it is clear from the curve that, the variations of the air-gap torque produced by the two machines do not vary significantly from each other. From this observation it can be deduced that all the quantities in the motor, such as stator current, do not vary significantly in the two machines when running in parallel.

7. CONCLUSIONS

When one synchronous motor fails, it may be

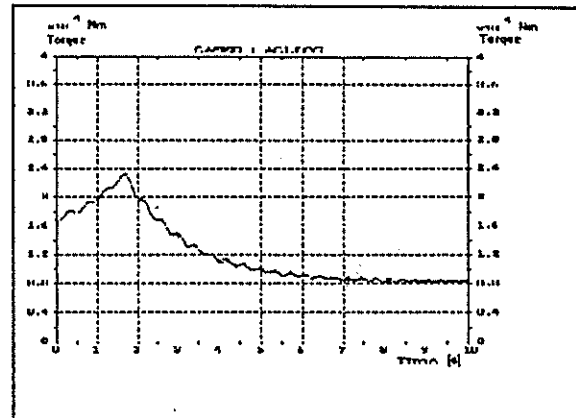


Figure 4 - Torque on synchronous motors during step increase in load on DC motor

necessary to replace it with a spare motor which is of the other design. When investigating the dynamic responses of the two types of motors when they are running independently, it is clear that the potential for problems may exist. However in order to fully appreciate the complete problem, the motors had to be simulated in a complete system.

From the simulation runs of the DC motors being started and a sudden torque step increase being applied to the motors, it was shown that the developed torques of the two different synchronous machines did not differ significantly, in fact the discrepancies were small. Further the differences in the two torques decayed, showing that the overall system is stable. These results show that the two motors should be able to be run in parallel, without modification to the controllers or the field resistance.

It must be born in mind that these are only simulation results, and that effects such as the saturation of the magnetising reactance and the rope dynamics have not been accounted for. Despite these limitations, the results show that the system is very stable. In order to be completely satisfied that the two different types of synchronous motors can be run in parallel, a test at site should be performed. Variables which should be measured so as to determine whether any problems exist on the actual machines are:

- Field current of each synchronous motor
- A line current of each motor
- The speed of each synchronous motor

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HIGH EFFICIENCY, LONG LIFE ACTUATOR FOR PV WATER PUMPING APPLICATIONS

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Abstract. A typical diaphragm configuration, photovoltaic (PV) water pumping system consists of a DC permanent magnet motor driving a diaphragm pump via a cam. The typical efficiency is low and its life span short. The use of a linear DC motor or actuator as drive is suggested. A higher efficiency and life expectancy could be achieved. The aims of the project were to measure the pump, evaluate possible electromagnetic configurations with a Finite Element Analysis (FEA) package and to evaluate different electromechanical options. The results obtained for these criteria will be discussed concluding with a choice of the best topology.

INTRODUCTION

A typical DC submersible diaphragm pump for small stand alone or hybrid PV water pumping systems comprises of a DC motor, a cam of some sort to convert the rotating to push-pull motion and a diaphragm pump section.

The efficiency of the complete diaphragm pump when utilising a converter (maximum power point tracker) is typically 15% with a 70 m head and a 94 Wp supply. The problems associated with these drives arise from cam losses, insufficient cooling at slow speeds and sideways forces on the rotor shaft. As a result, the bearings and brushes last for only 6 months to a year.

The suggested solution is to replace the motor and cam by a linear motor or actuator. By designing the plunger correctly a better efficiency and much longer life expectancy can be obtained. This paper will briefly discuss the measurements on the pump section, a detailed evaluation of possible electromagnetic configurations and results measured on the prototype.

CHARACTERISATION OF PUMP

The design parameters like thrust force and stroke length were measured on a commercial diaphragm pump and is shown in figure 1. The stroke length was measured as 3 mm. For a 30 m head and a displacement of 3 mm, a force of 375 N is theoretically required if a pump efficiency of 90% is assumed.

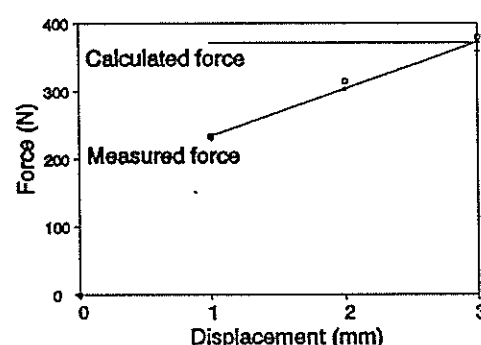


Figure 1: Force/displacement characteristics at 3 bar

ANALYSES OF ELECTROMAGNETIC ACTUATING CONFIGURATIONS

With the load characteristics of the pump known, possible actuating structures could be analysed. Because of cost considerations, it was decided to firstly search for a single plunger, reluctance type, linear motor.

a. Tapered Plunger

The first possibility was to make use of the idea that was optimised by Lequesne (1). The configuration was simplified and is shown in figures 2(a) and (b).

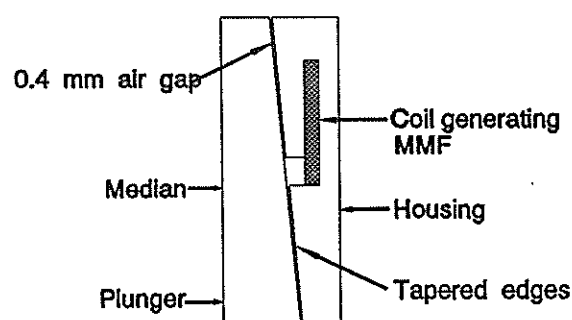


Figure 2(a): Tapered plunger configuration

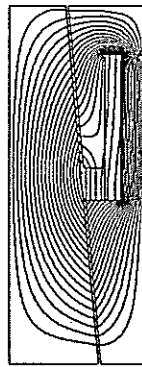


Figure 2(b): Tapered configuration with flux lines

The flux density in the air gap and the air gap length were chosen as 1.8 T and 0.4 mm respectively. The obtained force/displacement characteristic is shown in figure 3. Unfortunately, the plunger material severely saturated, making this option doubtful.

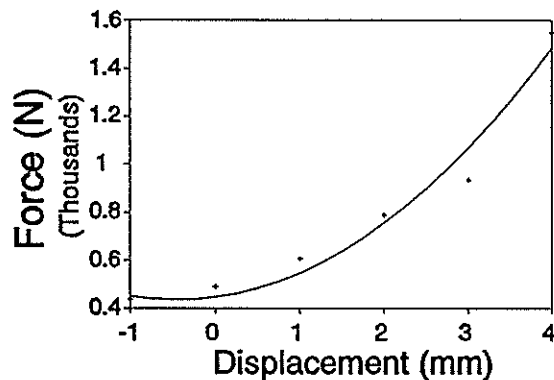


Figure 3: Simulated force/displacement results for the tapered plunger

The acceptability of this option will be evaluated later.

b. Square Motor Configuration

An attempt was made to apply the basic idea of Poloujadoff et al. (2), where longitudinal flux is used. The simulated configuration is shown in figure 4(a) and figure 4(b). Again the flux density in the air gap was 1.8 T with the gap length being 0.4 mm.

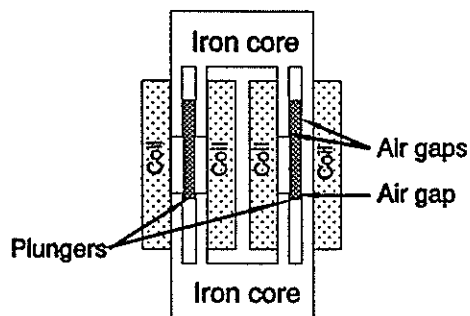


Figure 4(a): Square motor configuration

The configuration was simulated with a flux density of 1.8 T and a air gap of 0.4 mm and it was found that the two sliding pieces severely saturated.

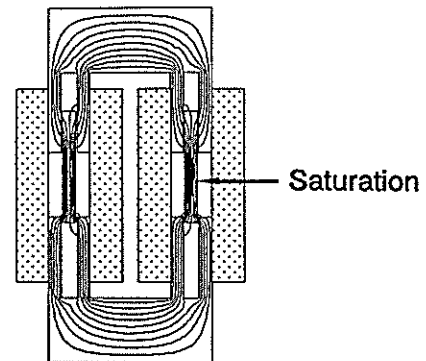


Figure 4(b): Square motor with flux lines

c. Cylindrical Plunger

A very promising solution was found when a cylindrical plunger inside a cylindrical housing was simulated. This is shown in figures 5(a) and (b).

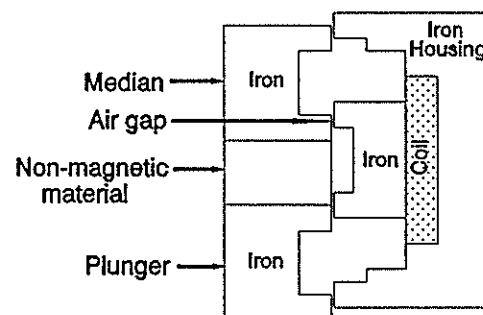


Figure 5(a): Cylindrical configuration

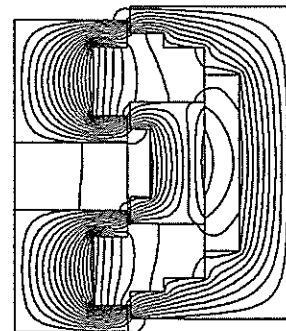


Figure 5(b): Cylindrical configuration with flux lines

From equation 1, the expected force during the stroke can be calculated.

$$F = \frac{B^2 \pi r g}{\mu_0} \quad (1)$$

where

r = radius of the plunger

g = gap length

μ_0 = permeability of air

The force calculated for the dimensions used, yielded a force of 108 N during the displacement. Again a flux density of 1.8 T and an air gap of 0.4 mm was assumed. The simulated result is shown in figure 6. The displacement at zero indicates the position where the opposing teeth start to align.

This configuration results in a fairly constant force/displacement characteristic over the required displacement. This is what is needed. If the amount of teeth increases, the developed force will also increase. There were no saturation in any of the parts either.

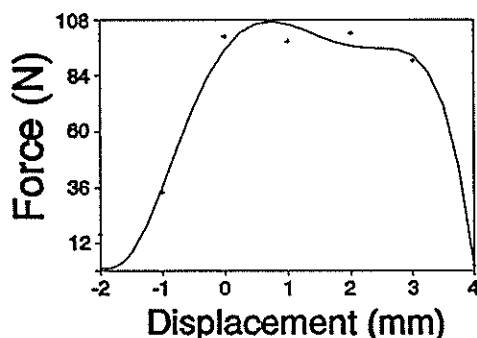


Figure 6: Simulated force/displacement characteristics for the cylindrical plunger

To prevent an upper frequency switching limit, an actuator capable of reciprocating was required. This will be most easily accomplished by adapting some of the basic ideas for using permanent magnets, as is being described in Binnie et al. (3).

d. Permanent Magnet Configuration

By incorporating the results of the cylindrical configuration and the basic ideas of (3), a prototype, as shown in figure 7(a), was designed with the magnets in a different position and the teeth designed to achieved the required force and stroke length. The PM denotes permanent magnet. In the resting position the flux lines are as is indicated in figure 7(b).

When a current is switched through the coil, the flux through one set of teeth are blocked and the actuator moves towards the other side. When the stroke is completed, the current is reversed, and the actuator moves in the opposite direction. With this configuration it is possible to obtain the desired force to pump water at a 30 m head. Because the configuration is basically that of the cylindrical case, the force/displacement characteristics has the same shape and the maximum expected force, namely 200 N, can be determined from equation 5. It should be remembered that this force is now only applied to one diaphragm pump, because of the use of the swivelling arm. In tabel 1, a comparison is made between the results obtained for and the characteristics of the different configurations.

Figure 7(a): Configuration of the

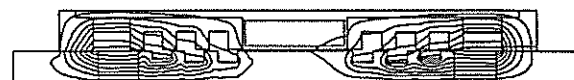
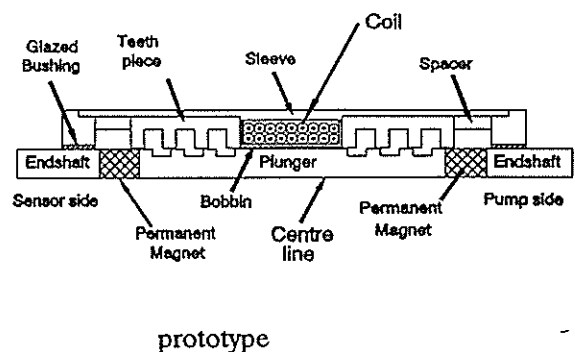


Figure 7(b): Prototype with flux lines

ELECTROMECHANICAL CONSIDERATIONS

Finally the mechanical side of the prototype was analysed. Having a single plunger, would ease construction, but it would be heavy, thus limiting the reciprocating frequency, and it will cause vibration. Because the diaphragm pump section consists of two small independent pumps, it is possible to have two independent actuators, one for each of the two pumps. However, calculations yielded that it was possible to develop more force with a coupled system, where there are two independent actuators, coupled by a pivoting arm, than in any of the two mentioned cases. Because the two plunger shafts are coupled, the two developed forces will add and that force is applied to only one diaphragm pump at a time.

RESULTS

A prototype was developed and measured as is shown in figure 8.

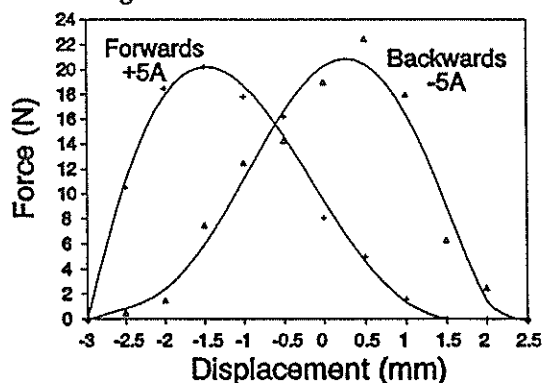


Figure 8: Measurements on the actuator

If the current is applied when the plunger is at the start of its stroke, the developed force increases, but after reaching a peak, it drops off again instead of staying constant. If the stroke is completed and

Table 1: Comparison of the different actuator configurations

	Advantages	Disadvantages
Tapered Configuration	1. simple construction 2. balanced radial forces - less friction	1. heavy plunger --big inertia will complicate early morning start-up 2. severe saturation of plunger 3. mechanical vibration 4. too long flux path
Square Motor Configuration	1. low plunger mass 2. flux path is long, but flux creates useful force several times 3. balanced sideways forces on plungers-less friction	1. not capable of delivering the required force without severe saturation of the plungers 2. very complicated construction 3. resultant force decreases with stroke
Cylindrical Configuration	1. balanced radial forces -less friction 2. capable of delivering the required force without saturating 3. flux is made to work several times	1. heavy plunger - high inertia 2. can only generate a force in one direction
Permanent Magnet Configuration	1. capable of delivering the required force at a 30 m head 2. plungers lightweight - low inertia - easier early morning start-up 3. balanced radial forces - less friction 4. low mechanical vibration 5. can deliver forces in both directions	1. more expensive option 2. small construction tolerances

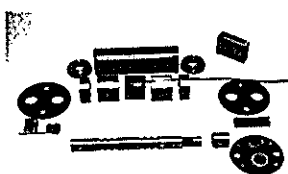
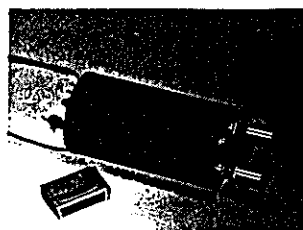
the current is reversed, a force is developed in the reverse direction. The magnets were tested and found to be not according to the factory specifications.

CONCLUSIONS

It has been established that the possibility exists to replace the DC motor and cam with a linear actuator solution without major system performance deterioration. Of all the simulated configurations, the two cylindrical linear actuators, coupled by a pivoting arm, posed the most promising solution. It was necessary to decrease the flux density in the air gap to 1.4 T and to increase the air gap length to 0.6 mm in order for the permanent magnet solution to be manufacturable. A prototype was developed and tested. In principle, the actuator seems to work fine, but further research in this field is necessary. Compared to the original product, the actuator solution shows great promise as far as the majority of the considerations are concerned. A better efficiency would be possible, because friction is greatly reduced. If these ideas could be incorporated into a spatially more optimal configuration, an increase in head would be possible.

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THYRISTOR CONTROLLED SERIES COMPENSATION OF INTERCONNECTING TRANSMISSION LINES

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ABSTRACT: Thyristor controlled Series Compensators (TCSCs) are attracting interest because of the opportunities they offer in the control of power flows through meshed circuits, their ability to modulate power flows and their potential to damp subsynchronous resonances. This paper examines the use of TCSCs in a typical interconnection configuration and their effectiveness in damping a low frequency inter-area oscillation.

Keywords - FACTS, Controlled Series Compensation, SVC, Interconnection, Modal damping

1. INTRODUCTION

Series compensated high voltage AC (HVAC) transmission lines are widely used to control flows and reduce reactive losses in transmission lines. The earlier problems associated with subsynchronous resonance are now well understood and can be avoided if proper design procedures are adhered to.

Adjustable series compensation in the form of a Thyristor Controlled Series Compensator (TCSC) is more flexible than fixed series compensation and offers, in addition to the power control functions, the benefits of stability improvement and reduced exposure to subsynchronous resonance (SSR) problems. TCSCs are a relatively new technology with three devices in service at present. The control bandwidth of the TCSCs is high enough to be considered for damping of lower frequency electromechanical oscillations such as might exist in a weak interconnection between two power systems.

Preliminary studies have been conducted into an HVAC interconnection between two power systems. The investigations included dynamic simulations of the system, formulation of suitable control strategies and the selection of appropriate controller input signals.

This paper describes the TCSC technologies investigated and their effectiveness in improving the damping of an inter-area oscillation in a simple test system.

2. Study system

Figure 1 shows the system used to illustrate the use of TCSCs in an HVAC interconnection. The test system is set up to be representative of two actual power systems and the simulation results are consistent with those obtained in more detailed studies [1,2].

The interconnection is formed between busbars C and D which have fault levels of around 1500 MVA and 5000 MVA respectively. Busbars B and C have a nominal

voltage of 330 kV while busbars D and E are 275 kV nominal. SVCs rated at -100 to +250 Mvar are located at either end of the inter-connection for voltage control and supplementary damping at times of low transfer.

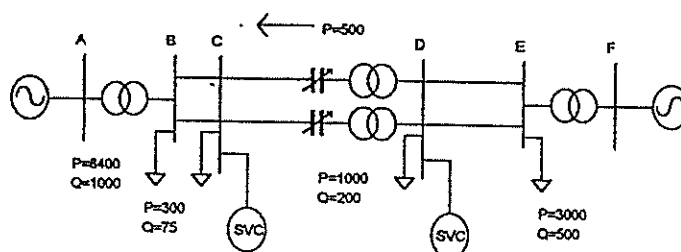


Figure 1: Test system with a 500 MW transfer on the interconnection from D to C

The system has a lightly damped inter-area mode with a frequency of approximately 0.6 Hz. For satisfactory performance of the interconnected system, supplementary damping of this mode would be required. In the above example, damping is provided using the TCSC and SVC devices.

3. Damping Effectiveness of the TCSC

The degree to which the damping of the inter-area mode can be improved depends on the quantum of energy that is used to modulate the tie line power flow. This quantum of energy is determined by the power flow through the TCSC and the dynamic range of the variable reactance. It is clear, for example, that at low tie line power flows, the effect of reactance modulation would be negligible and no damping improvement can be expected [1-3]. The quantum of energy available for modulation increases with power flow through the TCSC, making damping contributions at times of high tie line transfers significant [4].

4. Practical TCSC realisations

There are two types of TCSC currently in service: the thyristor switched series capacitor and the series capacitor with a thyristor controlled reactor (TCR) connected in parallel. The former provides discrete changes in series compensation while the latter can adjust the series compensation continuously over a given dynamic range. The systems will be discussed briefly.

4.1 Discrete TCSC

The discretely switched TCSC offers a number of advantages. The control system for the thyristor switch is simple since no firing angle delays are employed. This also means that no harmonics are generated as currents in the

valve or capacitor are always sinusoidal. Switching of the thyristors can be timed to minimise transients. The disturbances from capacitor insertion and removal could possibly cause problems if an SSR condition existed but would otherwise allow rapid switching to occur without serious effects to the rest of the system [5].

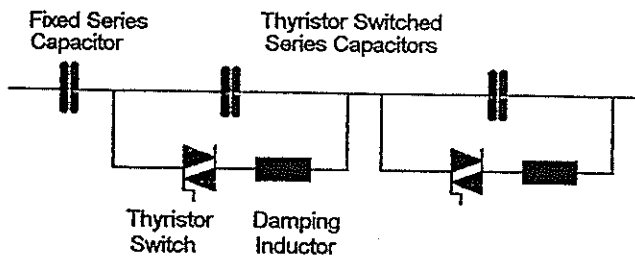


Figure 2: Thyristor switched series compensation

The switched TCSC is thus able to modulate the series reactance in a stepwise manner to provide supplementary damping of oscillation modes. At low oscillation amplitudes, however, the size of the discrete series element will determine the damping performance. Under very lightly damped conditions, the discrete changes in the inserted reactance may cause limit cycles to be established [6].

4.2 Continuously variable TCSC

Phase control of a TCR in parallel with a series capacitor element gives continuous control over the inserted reactance for a limited range. Typically, the TCR reactance will be smaller than that of the series capacitor. This means that the firing angle of the thyristors must be constrained to avoid the parallel resonant condition. The firing range is typically variable from 180° to about 145° , giving a variation in the reactance of the series element from 1 pu to about 4 pu. It is also possible to use a firing delay of 90° at which point the inserted reactance would be inductive. This mode of operation is referred to as Thyristor Switched Reactor (TSR) and provides the possibility of bang-bang control under large signal disturbances. [6,7]

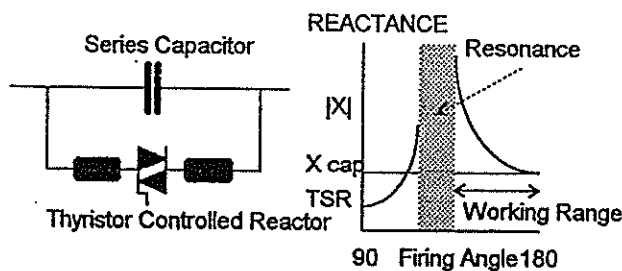


Figure 3: Continuously variable TCSC

The voltage across the series capacitor is a function of the current through the device and the delay angle of the TCR valve. The greater the firing angle delay, the higher the inserted reactance and the higher the voltage. The voltage withstand of the capacitor is time dependent, allowing operation at high voltages for short periods.

The non-linear currents through the TCR produce harmonic currents which circulate mainly in the capacitor. There can still be a significant harmonic injection into the system, particularly as the firing angle approaches the resonance point.

The continuous control of the series reactance requires more complicated thyristor control circuitry but allows damping improvement even for low level oscillations.

4.3 Surge arresters

Conventional series capacitors require parallel connected surge arresters of high energy rating to protect against high voltages caused by transmission system faults. The rating of these surge arresters can be greatly reduced where thyristors are connected in parallel with the series capacitor element as they can be fired to rapidly reduce the voltages developed across the capacitor.

For both of the TCSC types discussed, utilisation of the thyristor valve to reduce the energy rating of the capacitor surge arresters can reduce the cost differential between conventional and thyristor controlled series compensation.

4.4 Auxiliary power

As with conventional series compensation, the TCSC componentry is mounted on a high voltage platform. Transmittal of signals to and from ground based equipment for control, measurement and protection purposes makes extensive use of fibre optic systems. Auxiliary power on the platform to drive control and measurement devices poses a further problem. One option is to use current transformers connected in series with the TCSC device so that, provided currents are above a certain threshold, energy may be extracted from the through current. It is normal under conditions of low power flow to bypass the series capacitor bank. This is clearly not desirable for a TCSC in a tie line where the power swings could cause the currents to cross the threshold several times during a disturbance. Even if auxiliary power could be supplied from another source, say a voltage transformer, arrangements would have to be made to ensure correct firing of the thyristor valve during the phase reversal of the current.

The use of optically fired thyristors overcomes many of these problems provided the control and measuring equipment can either use optically supplied energy [5] or other control arrangements can be employed to safely fire the valve under low current conditions.

In applications where transfers in the TCSC equipped transmission line are normally high enough to provide the auxiliary energy, the CT method is quite acceptable. Short term storage on the platform could also provide sufficient auxiliary power to cater for rapid swings through zero (or low) current conditions. The recovery of the auxiliary power circuitry following a current zero can be achieved within a few cycles of power frequency.

4.5 Subsynchronous resonance

The switched TCSC faces the same problems as conventional fixed series compensation but has the ability to change the level of compensation should an SSR condition be detected. There is also some concern that the transients associated with the switching of a series element could excite oscillations between the shaft systems and the network.

The continuously controllable TCSC is less susceptible to SSR because the impedance characteristic at SSR frequencies tends to be inductive. The impedance of the

parallel combination of a TCR and a capacitor is not linearly dependent on frequency, as is the case with fixed series capacitors. The inductive impedance avoids resonant conditions between the network inductance and the TCSC.

Although not demonstrated yet, it is conceptually possible for a control algorithm to be developed which would allow the continuously controlled TCSC to actively damp an SSR condition.

4.6 Selection of the TCSC input signal

In the case of a TCSC on the interconnecting transmission circuits, the inter-area mode would be easily identifiable in the tie line power flow signal. If the modulation controls are based on this signal, the action of the TCSC could be made to directly affect the participants in the mode, leading to effective supplementary damping. The simple control structure used for this investigation is given in Figure 4.

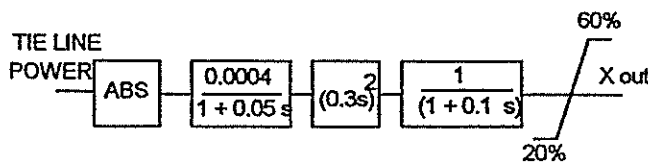


Figure 4: Controller structure for the TCSC

The dynamic range of the TCSC is based on percentage compensation of the 5.7% impedance circuits between busbars C and D in Figure 1. The absolute of the power signal is required because the control action is the same irrespective of whether the relative machine angles are separating in a positive or negative sense.

5 Static Var Compensators

Apart from the normal voltage control functions, SVCs are able to provide supplementary damping for electromechanical oscillation modes by virtue of demand modulation. A stabilising signal can cause an SVC to vary the voltage in an area. Since loads have a voltage dependency, this will cause both the active and reactive components of the load to change in sympathy. A simple load model is shown below:

$$P = P_0 \left(\frac{V}{V_0} \right)^\alpha \quad \text{Where typically } 1 \leq \alpha \leq 2 \quad (1)$$

$$Q = Q_0 \left(\frac{V}{V_0} \right)^\beta \quad \text{Where typically } 2 \leq \beta \leq 4$$

If the demand in an area is modulated with an appropriate phase relationship to the a selected oscillation mode, damping torques can be produced. The extent of damping achievable in this manner depends on the size of the SVC, the tolerable voltage variations, the area load and the voltage dependency of this load. The active component of the load, which typically has an aggregate response somewhere between constant current and constant admittance, is normally more significant than the reactive load for purposes of supplementary damping. Lower voltage dependencies, for example where $\alpha < 1$ in (1), would reduce the effectiveness of an SVC in providing supplementary damping.

The stabilising signal selected for the SVC was tie-line

power flow, as in the TCSC, as this is available at the SVC location and provides identification of the inter-area mode.

6. Test system performance

Figure 5 gives the simulated response of the test system to a fault at busbar D followed by tripping of the tie lines. This is not the most severe fault from a transient stability viewpoint but serves to illustrate the damping of the inter-area tie mode. The transfer conditions are as indicated in Figure 1 with a 500 MW export from busbar D to C. Conservative assumptions were made for the voltage dependency of loads in that the active component was assumed to vary with the square root of the voltage change ($\alpha=0.5$). This has the effect of reducing the damping of the inter-area mode [1,2]. The improved performance with the TCSC in service is apparent.

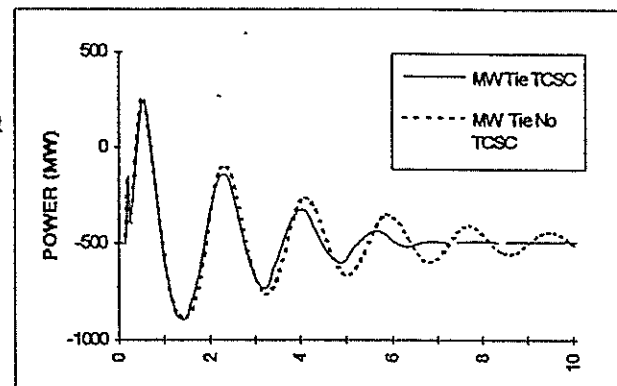


Figure 5: Simulated responses for 500 MW export from busbar D to C.

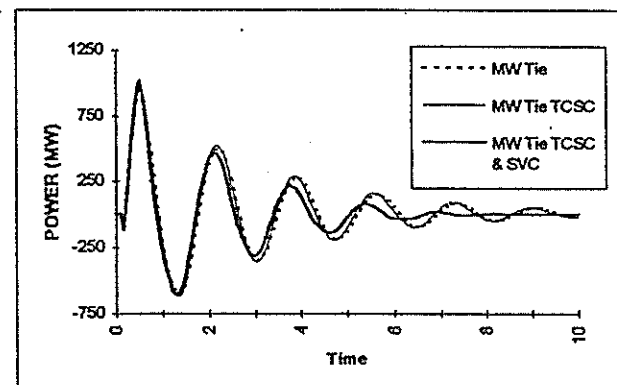


Figure 6: Simulated responses for zero transfer between busbars D and C.

At times of low or zero transfer, the TCSC is virtually ineffective and the SVCs have to provide the supplementary damping. The performance at time of zero transfer is given in Figure 6.

The effectiveness of the SVCs is demonstrated in the zero transfer case. Figure 7 shows the simulated response for the conditions of Figure 5 but with the SVCs and TCSCs in service.

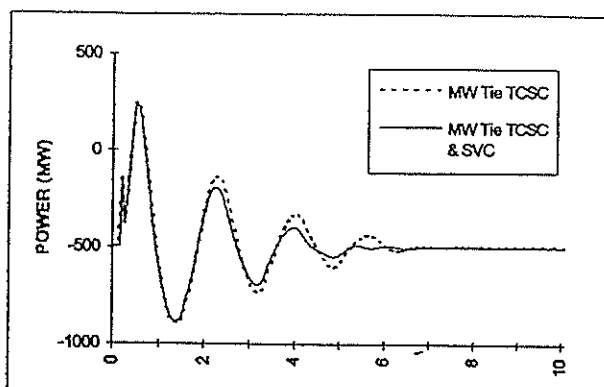


Figure 7: Damping improvement with both TCSCs and SVCs for 500 MW export from busbar D to C.

Figure 8 gives the response for a 500 MW transfer from busbar C to D. This is the most severe of the three operating conditions examined as the transmission circuits from B to C, which have a combined impedance of around 6%, become more heavily loaded. For transfers in the reverse direction, this part of the transmission system becomes less loaded and more stable performance could be expected and is evident in Figures 5 and 7. In Figure 8 it is apparent that the damping of the low frequency inter-area mode has improved but that a higher frequency mode is excited towards the end of the plot. This is an interaction between the controller of the SVC at E and the TCSC. The controller for the TCSC has a fairly basic structure, as shown in Figure 4, and the given controller, although having the correct gain and phase at the inter-area modal frequency, has an unacceptably high gain for higher frequencies. A controller structure with more flexibility would allow for a faster gain roll-off after the inter-area modal frequency and should be able to reduce the interaction.

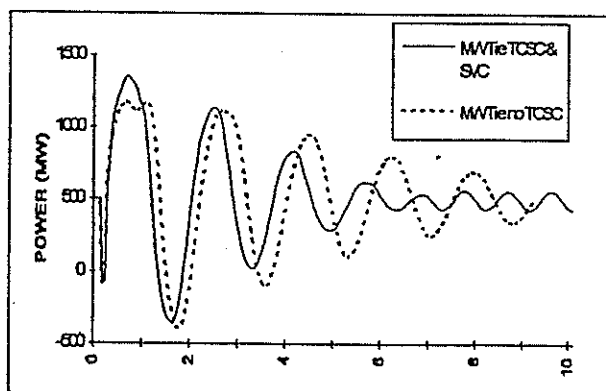


Figure 8: Simulated responses with no supplementary damping and with both TCSCs and SVCs for 500 MW import from busbar C to D.

7. Conclusions

The damping improvement in the test system shows that even for moderate levels of transfer and a relatively small dynamic range on the TCSC, significant improvements in damping can be realised. At times of low transfer, however, a TCSC can not provide supplementary damping. Problems with auxiliary power can also affect performance of a TCSC at such times and possibly even for power swings which approach zero or reverse flow direction.

The two types of TCSC currently being promoted both have

benefits and disadvantages. For large power swings, the two technologies give similar performance if constrained to the same dynamic range. If TSR operation is modelled on the continuous TCSC, improved performance would be expected. Before selecting a particular technology, it would be prudent to examine the effects of the TCSC on both harmonic levels and possible SSR conditions.

Although not discussed in the text, the relative costs of TCSC and SVC provided damping make TCSCs an attractive proposition, especially where series compensation would have been installed anyway. TCSCs have a damping effectiveness which increases with transfer while SVCs would be dependent on the loads available for modulation. As the loads vary with time, studies encompassing all expected loading conditions would be required to ensure satisfactory damping enhancement.

The effectiveness of the TCSC in an interconnection of the type studied would depend on the expected utilisation of the tie, the times for which the flows would be low or zero and the manner in which the damping of the inter-area mode changes with transfer.

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INITIAL INVESTIGATIONS INTO CONVERTOR BASED SERIES COMPENSATION

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ABSTRACT

This paper reports on the results of a preliminary study into inverter based series compensation for Flexible AC Transmission Systems (FACTS). Simulated results of a proposed system are presented to verify the control concepts, and three different levels of inverter modelling are then considered. The results indicate that a simplified model can accurately represent the effect of a FACTS device on the transmission line currents in power system transient studies.

INTRODUCTION

Power utilities are encountering increased difficulty in building new high voltage transmission lines for environmental and economic reasons. Due to stability criteria brought about largely through the slow methods used to control power systems, existing transmission lines are generally loaded only up to their surge impedance limit. This surge impedance limit is considerably less than the theoretical power transfer limit of a line due to thermal considerations, particularly in the case of high voltage lines. This has led to research into FACTS (Flexible AC Transmission Systems) technology which aims to make the power system more dynamically controllable by means of high power electronic devices, thus enabling the system to be safely operated nearer to its theoretical limit.

Typical FACTS devices include power electronic controlled phase shifters, series compensators, dynamic brakes and advanced static VAR compensators. Many of these devices are not new and indeed have been operational in power systems for some time in mechanical form. The capability of FACTS technology to achieve high speed continuous control of these devices however, offers two scenarios: either an increased power transfer over existing lines for the same stability margin (thus removing the need for new lines), or the improvement of the stability margin for problem lines whilst maintaining current power transfer levels. The reader interested in a detailed treatment of FACTS and its objectives is referred to Refs. [1] and [2], but for the purposes of this paper it is sufficient to state that the objective of a FACTS device is to provide high speed control of a parameter which affects the power transfer capability of a power system.

FACTS SERIES COMPENSATORS

One of the FACTS areas which shows most promise is that of controllable series capacitance. The inductance of long transmission lines limits their steady state power transfer capability, and the inclusion of fixed capacitor banks in series with these lines has the effect of cancelling out part of this inductance, making the line appear electrically shorter. Such fixed series compensation has been used in power systems worldwide for many years. More recently it has been shown that in systems where concerns for stability limit the power transfer, if a small amount of the series capacitance (as little as 10%) is made *continuously controllable* through a FACTS

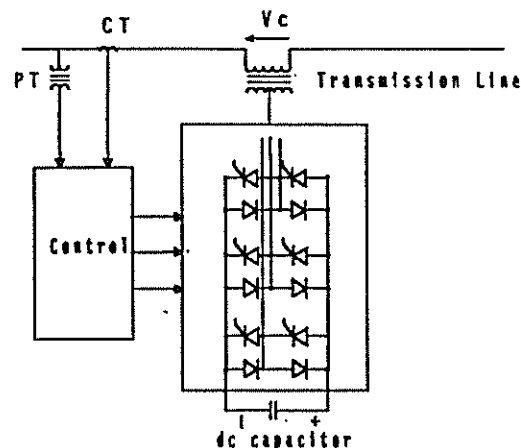


Fig. 1 Series compensator based on voltage source inverter

device, this power transfer limit can be dramatically improved [3].

The extent to which series compensation can be introduced to the power system is often limited by the very real threat of subsynchronous resonance (SSR) [8]. Controllable series compensation however has the *potential* to avoid SSR or dampen SSR from other sources. However, for this potential to be realised requires thorough understanding not only of how the capacitive elements of these new devices will affect power system resonances but also of how their control systems will interact with these resonances. The first step towards such an understanding is the development of reliable models of these FACTS devices for inclusion in the simulation programs used to study the behaviour of the power system dynamics.

SERIES COMPENSATION BY VOLTAGE SOURCE INVERTER

Gyugyi [4] first proposed the concept of a GTO-based solid state series capacitive compensator, using a voltage source inverter as seen in Fig. 1. The inverter is used to inject a voltage V_c in series with the line such that this voltage V_c lags the line current by 90 degrees. The amplitude of the voltage V_c is controlled in such a way that it is proportional to the amplitude of the line current. Mathematically this can be represented by the phasor equation

$$V_c = k I e^{-j\frac{\pi}{2}} \quad (1)$$

such that the voltage injected by the inverter appears to the transmission line to be a capacitive reactance. Furthermore, by changing the constant of proportionality k , the degree of compensation can be continuously controlled. The solid state series compensator also has the potential to avoid SSR, since with the correct control action the inverter injects only a voltage at the fundamental frequency and presents

theoretically zero impedance at all other frequencies. The success of the scheme however depends upon the control strategy adopted to achieve the behaviour in eqn. (1).

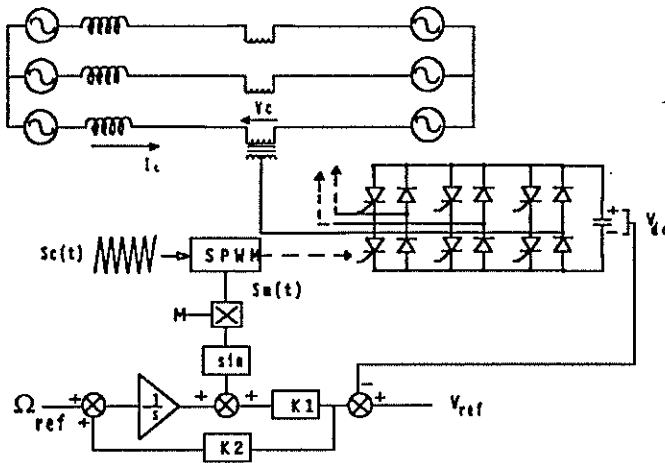


Fig. 2 Proposed control strategy to achieve reactance property

One such strategy to achieve the desired control of the inverter in Fig. 1 has been proposed by Ooi et al [5] and is shown diagrammatically in Fig. 2. In order to outline the operation of the scheme in Fig. 2, let us assume first that a voltage exists on the dc link capacitor C_{dc} . The ac voltage waveform which then appears at the output of the inverter is achieved using the sinusoidal pulse width modulation (SPWM) technique, by comparing the output of the $\sin$ function block with a triangular carrier wave of sufficiently high frequency, to generate the switching logic for the devices in the inverter. In this way the fundamental component of the ac voltage waveform at the terminals of the inverter follows the output of the $\sin$ function block in frequency and phase, and its magnitude is determined by the dc link voltage and any multiplying factor that is supplied at point M.

Now that the technique of injecting the ac voltage has been established, consider again the dc link voltage; if the ac voltages and currents at the injection transformer primaries happen to be in phase, then the inverter is accepting power from the line and the capacitor voltage must thus increase. Likewise if these transformer voltages and currents happen to be out of phase, then the line is accepting power from the inverter and the capacitor voltage must decrease. However, the dc link voltage regulator feedback loop controls the capacitor voltage to be a set value V_{ref} by steering the frequency and phase of the injected ac voltage. When the error $\varepsilon = V_{ref} - V_{dc}$ in this voltage feedback loop is zero, no power is flowing in or out of the dc link to either charge or discharge the capacitor. This then implies that at the terminals of each series injection transformer the voltages and currents obey the relationship $\text{Real}(\underline{V}\underline{I}^*) = 0$, thus providing the required behaviour of a purely reactive impedance. The reactance is made capacitive by the correct choice of polarity in magnitude control M.

The sine wave template passed to the SPWM control is steered by the outputs of the voltage regulator loop, and can be described mathematically as

$$S_m(t) = M \sin[(\Omega_{ref} + K_2 \varepsilon)t + K_1 \varepsilon] \quad (2)$$

where $\varepsilon = V_{ref} - V_{dc}$

so that when the error ε is nulled the sinusoidal output is of the required frequency and phase to yield the steady state reactive property. A detailed discussion of this control concept can be found in Refs. [5-7].

MODELLING THE COMPENSATOR FOR SIMULATION

In order to assess a system such as the one proposed in Fig. 2 it is necessary to develop reliable models, which are able to predict the system's behaviour with sufficient accuracy, for inclusion in the large power system models that are typically used to study power system stability problems. Several simulation studies have been performed to this end using the Electromagnetic Transients Program (EMTP). For the initial studies presented here, the sending and receiving ends in Fig. 2 are represented as ideal voltage sources. The initial objective is to gain an understanding of the behaviour of the converter itself under the control scheme, and to determine the level of complexity of inverter modelling that is required to accurately represent this behaviour.

The first model considered is referred to as the Detailed Model. In this Detailed Model the GTO valves and freewheeling diodes are represented in the EMTP as ideal switches which can be opened or closed by means of control variables in the program. The correct switching action of the GTO's and diodes is ensured by monitoring both the PWM switching commands and the node voltages in the inverter. Fig. 3 shows the results of a simulation of the system of Fig. 2 employing the Detailed Model of the inverter. The initial conditions in the simulation are specified such that the dc capacitor is pre-charged to its reference voltage of $V_{dc} = 100$ V. The first graph in Fig. 3 shows the error signal ε representing the difference between the reference voltage and the actual dc link voltage.

The error ε is initially zero by virtue of the initial condition imposed on the capacitor and not due to the correct phase relationship of the injected voltage, and so it immediately begins to increase as the actual phase angle of the injected voltage causes the capacitor to discharge. This error then begins to steer the injected voltage vector to its correct position with respect to the line current vector. This is done in two ways: the gain K_1 results in the phase of the injected voltage being advanced by $K_1 \varepsilon$, (which with respect to the angle error is proportional control) whilst the gain K_2 results in the angular speed of the injected voltage being increased by $K_2 \varepsilon$ (which with respect to the angle error is integral control). Fig. 3 shows that the error ε converges to approximately zero at steady state.

The bottom graph in Fig. 3 shows the injected voltage and the transmission line current of phase a during the first 0.15 seconds of the simulation. The voltage waveform is seen as the typical SPWM pulse pattern since no harmonic filters have been included across the transformers. The pulses appear to be irregular since due to space considerations the

simulation output is plotted at large time intervals. The line current (which has been magnified by 10 for ease of comparison) is approximately sinusoidal due to the smoothing effect of the line inductance.

Comparing the top and bottom plots, the effect of the control action can readily be seen. At the start of the simulation the voltages and currents are almost exactly out of phase, resulting in discharging of the dc link capacitor and hence the rapid increase in the error ϵ . Within the first cycle the control has brought the voltage and current approximately into quadrature as the error again approaches zero. After 0.15 seconds the system has reached a steady state where the error is approximately zero and the line current leads the injected voltage by 90 degrees.

The simulation in Fig. 3 is a close representation of the system in Fig. 2, since the switching behaviour of the inverter is represented in detail. However its accuracy relies on the choice of a sufficiently small time step of simulation in the EMTP, since switching events can only occur in the simulation at fixed time intervals. Indeed, several runs of the case study in Fig. 3 at increasing time steps have indicated that the simulated predictions of the control behaviour become affected by step size at far smaller time steps than are required simply for numerical stability. In order to determine a reasonable step size for the Detailed Model of the system, the approach adopted has thus been to decrease the simulation time step until the results of successive simulation runs are unchanged. With this approach it has been found that for sufficient accuracy of the results, a single simulation run such as that of Fig. 3 requires several hours of computing time. Clearly this level of detail in modelling the system is prohibitive for power system studies which generally require longer simulations and the inclusion of additional high order models to represent machine dynamics. The detailed model of the inverter used in Fig. 3 is valuable however, since it gives insight into the practical aspects of the system necessary for constructing a working laboratory prototype. However for longer simulation runs a simplified representation is necessary.

For this reason two levels of simplification to the detailed model of the inverter are considered. The EMTP has a facility which allows the injection of a voltage in series with the transmission line whose magnitude at any instant in the simulation can be directly specified by control variables. This feature is used to model the effect of the inverter in Fig. 2 as a variable voltage source in the transmission line, by assuming the inverter instantaneously produces at its terminals the frequency and phase of the voltage demanded by the SPWM control. This means the EMTP does not actually model the changing inverter topology using switches which need to be monitored at every time step. However the magnitude of the ac voltage of the inverter is dependent on the dc link voltage, so the new voltage source in the EMTP includes a differential equation describing the charging of the dc capacitor which is solved at each step before calculating the injected ac voltage.

The first level of simplification is aimed simply at removing the burden of modelling the opening and closing action of the inverter switches in the EMTP. Thus at each step in the simulation after calculating the dc capacitor voltage, the SPWM determines the required switch states by comparing the template generated by the control signals [eqn. (2)] with the carrier wave (as though the inverter was being modelled

in detail). However the value of the injected voltage source in each phase is then set directly to be either the dc link voltage or zero according to the switch states. This Intermediate Model removes the changing network topology from the EMTP solution, but still models the ac voltages at the inverter terminals accurately, and thus still requires reasonably small time steps.

The second level of simplification, referred to as the Simplified Model, does not consider the switches in the inverter at all. In this model, the dc capacitor voltage is determined at each time step and the resulting control signals used to determine the template for the SPWM as in the Intermediate Model. However the template waveform is injected in each phase directly without determining the switch states. This model thus ignores the switched nature of the inverter voltages by assuming that only the fundamental of the desired ac waveform appears at the inverter terminals. This then removes the need for a simulation time step which is small in comparison to the time period of the carrier wave, and hence results in large savings in computer time.

Fig. 4 shows a comparison of the results obtained using the two levels of simplification described above, for the same case study as shown in Fig. 3. In the two left hand plots the results obtained using the Intermediate Model are compared directly with those from the Detailed Model study already shown in Fig. 3. The right hand plots show a similar comparison between the Simplified Model and the Detailed Model. In Fig. 4 the curves showing the behaviour of the error ϵ are almost identical in the case of the Detailed and Intermediate models. In the case of the Simplified Model the overall trajectory of the error ϵ corresponds to that of the Detailed Model, however the higher frequency ripples present in this error signal are not present in the Simplified model. This is as a consequence of the fact that the injected voltage in the Simplified Model only represents the fundamental frequency of the inverter voltage. This can readily be seen in Fig. 5 where the voltages and currents at the ac terminals of the inverter are shown for the Intermediate and Simplified models.

The high frequency oscillations superimposed on the error signal ϵ in the Detailed and Intermediate models in Fig. 4 arise due to the fact that the harmonics in the inverter voltage contribute to the charging and discharging of the dc capacitor when they interact with the line currents. Although harmonics are ignored in the Simplified Model, nonetheless the overall agreement between the error ϵ in the Detailed and Simplified models is sufficiently close. This is further evidenced by studying the transmission line currents in Fig. 4. The curves show that despite the simplifying assumptions made in the representation of the inverter voltages, the transmission line currents of all three models are almost identical. The transmission line currents are the link through which the device and its controller will interact with the other elements of the power system, and thus it is of most importance that any model considered accurately represents the effect of the device on these currents. Fig. 4 indicates that both the Intermediate and Simplified Models represent the effect of the inverter and its control on the line currents to almost the same degree of accuracy as the Detailed Model.

DISCUSSION

The results of a preliminary study into inverter based series compensation for FACTS have been presented. Since the

initial goal is to develop a sound understanding of the behaviour of the inverter under the proposed control, and to establish suitable models, the studies have been limited to a single line system fed from ideal voltage sources. The studies have verified the concepts of a proposed method of control of an inverter for such an application using EMTP simulations. Three representations of the inverter in the EMTP have been considered with varying degrees of complexity and the results compared.

All three representations of the inverter show close agreement when considering the interaction with the transmission line currents, but demand different amounts of computing time for sufficient accuracy. The Detailed Model requires several hours of simulation time in the EMTP even for the simple power system considered. The Intermediate Model results in a degree of time saving by avoiding topology changes in the network, but the time step requirements are the same as those of the Detailed Model. The Simplified Model results in a considerable time saving by representing only the fundamental voltage at the ac terminals of the inverter. The time step can be made more than an order of magnitude larger than required for the Detailed Model thus making the Simplified Model suitable for use in larger scale power system simulations.

Based upon the results presented it is felt that the Simplified Model is a reasonable representation of the inverter for this FACTS application. However further comparisons in which machine dynamics are included in the power system model are necessary to confirm this finding. The project is in its early stages, but the long term aim is to assess the impact of such FACTS devices on the SSR phenomenon in the Micro-Machines Laboratory at Natal University; as such the final judgement of the models will depend on their ability to correctly predict the behaviour of this laboratory test system.

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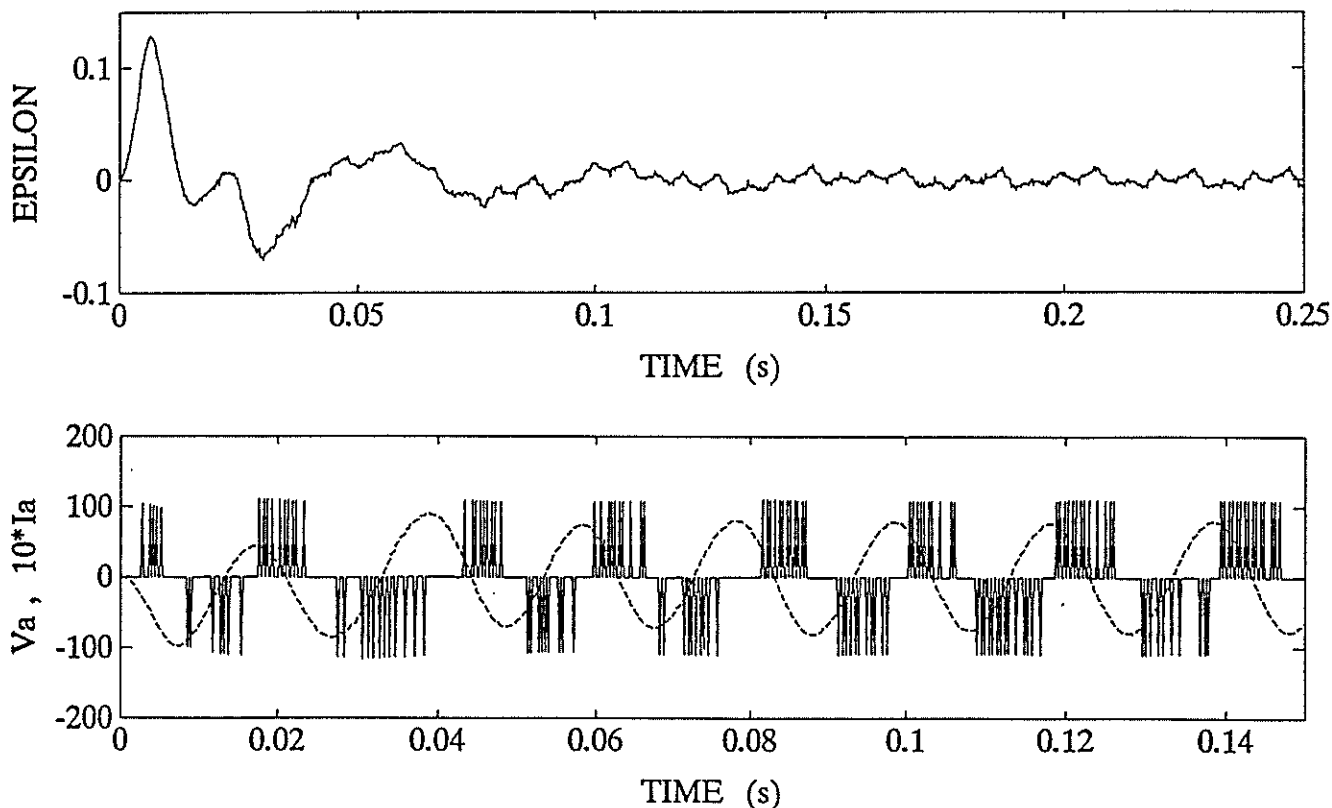


Fig. 3 Simulated results with Detailed Model of the inverter

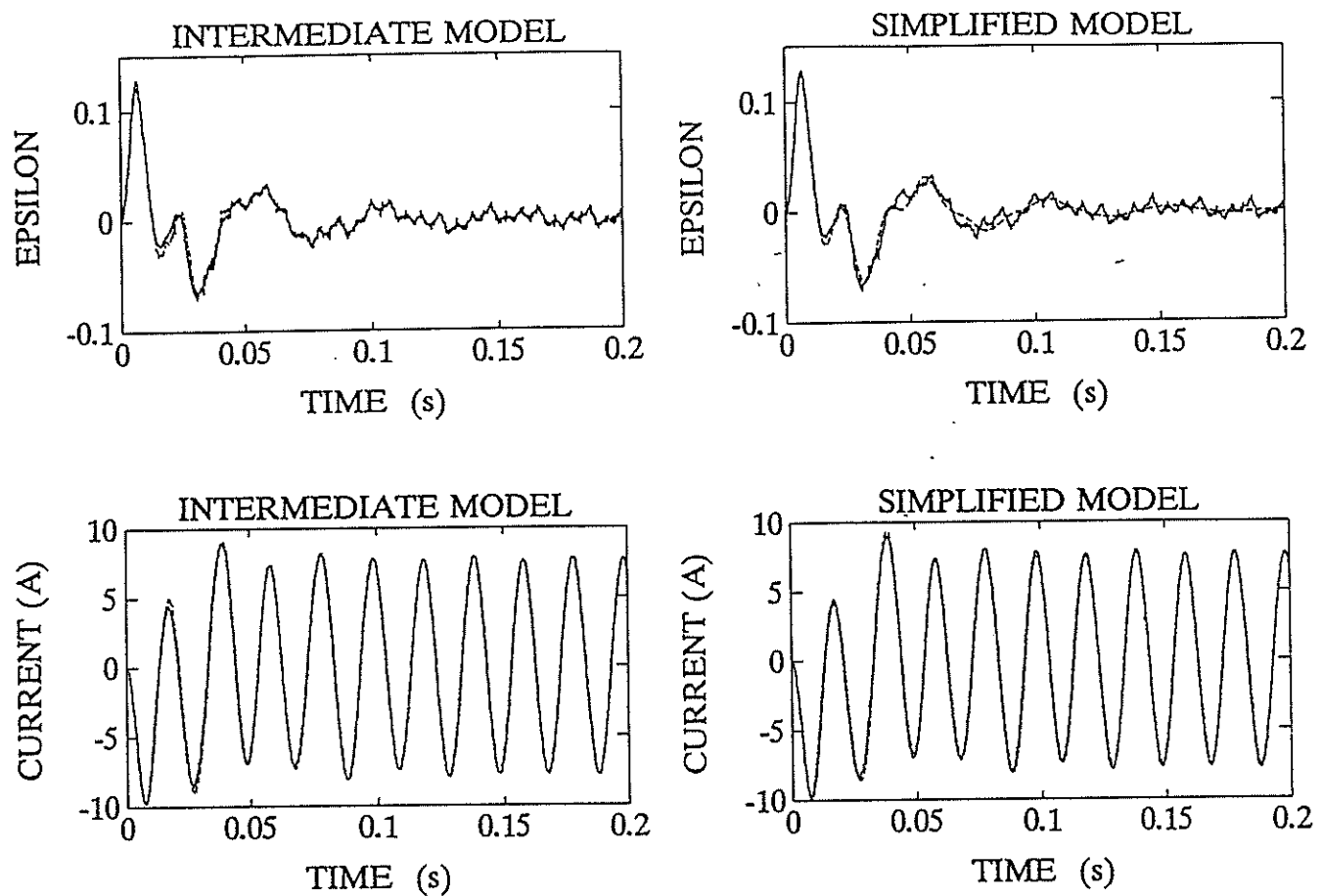


Fig. 4 Comparison of the results of Fig. 3 with the results of both the Intermediate Model and the Simplified Model

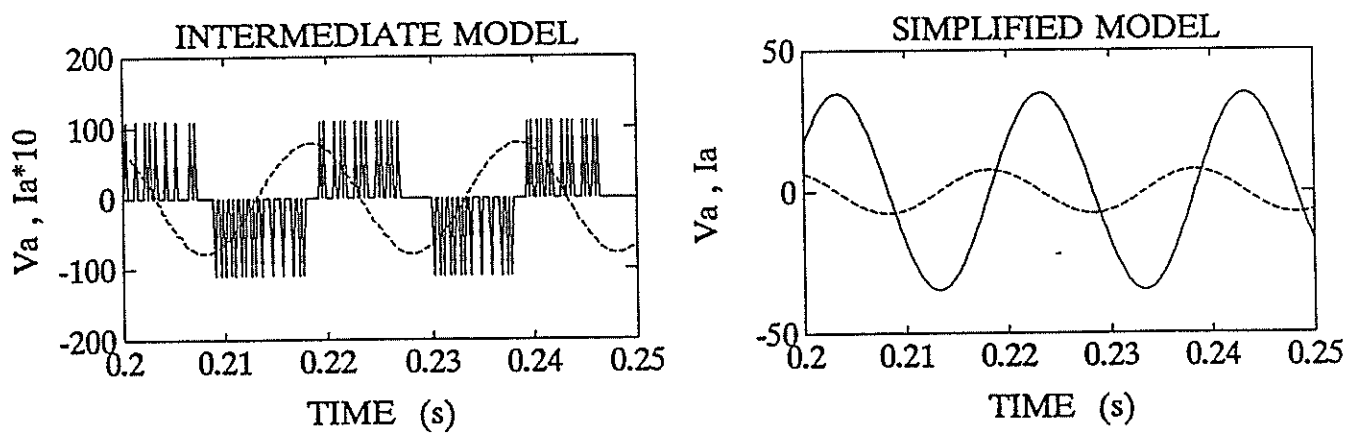


Fig. 5 Voltages and currents at the inverter terminals for Intermediate and Simplified Models

REVIEW AND PROPOSED SINGLE PHASE TO THREE PHASE CONVERTER

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SAUPEC Jan 12,13 1994

Abstract

In this paper existing single phase to three phase converter techniques are reviewed. A suggested method using a reduced size PWM phased locked single phase converter and transformer to form part of the required additional balanced two phases is investigated.

1 Introduction

Rapid extension of the electricity grid is crucial for the welfare and future development of South Africa. Three-phase power lines to rural, agricultural sector and light industries is not often economical. Consequently many areas are and will be supplied from single-phase mains. Single phase supply is adequate for most domestic appliance, however there are farms, hospitals, workshops and home industries where three phase motors and rectifiers are needed. The reason is that three phase equipment are generally more efficient, less expensive, more readily available, and more reliable than equivalent single phase equivalent. Additionally less demand is made from the supply if the single phase is first converted by some means to three phase and three phase equipment used instead.

Single phase motors have inferior characteristics such as low starting torque, very high starting currents, and pulsating torques which leads to mechanical failure and noise. Single phase transmission in the above ap-

plications are usually of limited ratings and therefore subject to severe voltage fluctuations. In addition three phase supply enables the use of a simple star-delta starters which reduces further by a factor of three the starting or "inrush" currents for loads such as pumps and fans. Moreover, the inverter drive of variable speed drives for induction motors need a D.C. link.[2] A single phase rectifier produces two pulses as opposed to six pulses per cycle for a three phase rectifier. Therefore a two-pulse bridge rectifier circuit leads to relatively high current pulses (crest factor) being drawn from the supply, hence generating a large amount of undesirable harmonics. The total harmonic distortion (TDH) of the single phase rectifier current is in the order of 45% as opposed to 5% for a three phase rectifier. The power factor of the former is about .75 as opposed to .95 for the later. Several single phase to three phase converters exist and have been proposed each have various economical and performances trade-offs. A few of these are now reviewed so as to compare with the one currently being investigated.

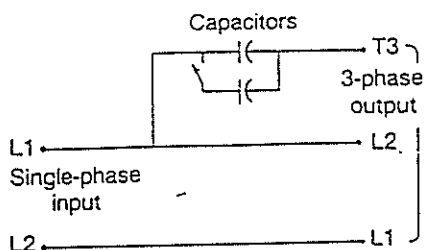


Figure 1: Capacitor phase-shift static converter

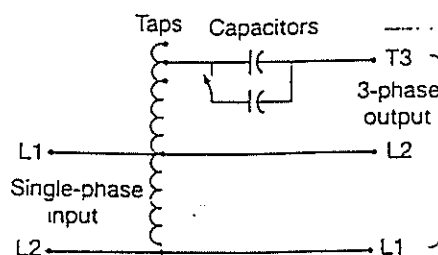


Figure 2: Autotransformer phase-shift capacitor converter

2 CONVERTERS REVIEW

2.1 Capacitor phase-shift Converter

A crude but sometimes effective method of running a three phase motor is to use an appropriate capacitor as shown in Fig 1. [1] The capacitor-type converter has no way of adjusting output voltage, or phase angle. Thus, it will always have some phase current imbalance. The motor hence must be substantially derated. Furthermore the line to line voltage would be the actual single phase line to neutral voltage. Three phase motors are normally available to run in delta. Hence the above converter would require a lower voltage motor. For example the usual 380V line to line voltage motor meant to start in star and run in delta would not be suitable unless one was to run underfluxed and at very light load. The voltages could of course be stepped up with a 3 phase transformer. A better converter is the autotransformer capacitor converter shown in fig 2. An autotransformer/capacitor converter usually offers the best performance for maintaining voltage and current balance. But the balance can only be maintained at a specific load. However more sophisticated types can maintain balance over a reasonable load range by adjusting the capacitance and autotransformer voltage this is achieved by tap changing and static semiconductor capacitor

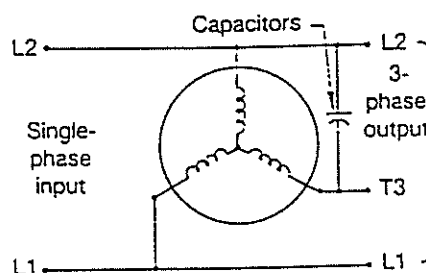


Figure 3: Rotary converter

switching. [3]

2.2 Rotary Converter

A rotary phase converter is an induction machine that operates on a single-phase supply and produces a three-phase output [4]. This method is known as the Ferraris-Arno system and has been widely used since its invention since the turn of the century. It has a stator frame with a symmetrical 3-phase winding and a specially modified squirrel-cage rotor. A large capacitor bank is placed across a coil group between one input line and the manufactured phase. It has now become prohibitively expensive if they are rated to supply direct on-line starting or pole-changing motor inrush currents. [5]

2.3 DC Link Converters

The more conventional and commercially available semi-conductor phase converter is a version of the DC link variable voltage and frequency converter. In this converter, a single-phase

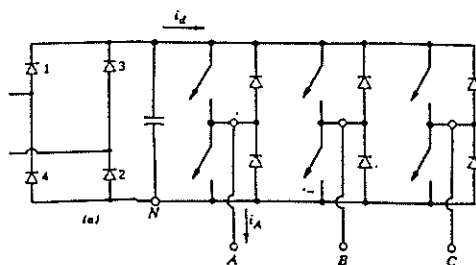


Figure 4: DC link converter

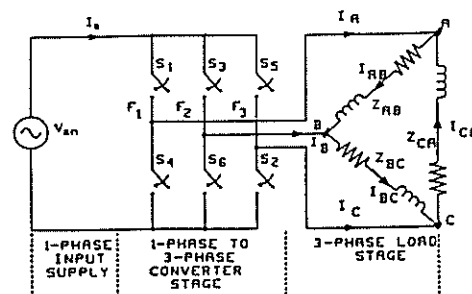


Figure 6: Single phase Cycloconverter

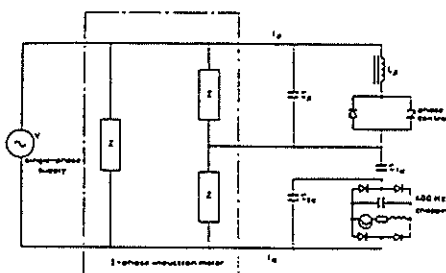


Figure 5: Electronically controlled reactor converter

rectifier is employed at the input to rectify the AC mains voltage. A three-phase inverter is employed at the output to convert the DC to a set of three-phase voltages. These converter are relatively expensive and availability of commercial units are limited to about 3 KVA. The output voltage is also low and therefore requires a non standard voltage motor. (e.g 230V line to line instead of 400V which is the norm in S.Africa) See Fig 4.

2.4 Reactive Phase Converter

This method (See Fig 5) uses two electronically controlled reactors so as to maintain balance under transient change in load. It is relatively simple and gives satisfactory operation of a direct-on-line starting [5], cage induction motor. However a low voltage non standard motor has to be used and start-delta start is not possible.

2.5 Cycloconverter Converter

This approach utilizes direct cycloconverter principles, [6] and its implementation requires the use of solid state switches only. See Fig 6 This proposed converter does not require a rectifier as the previous. However, it also has some disadvantages such as high third order harmonics and non standard low voltage of 55% of voltage input. This problem can be solved by the use of a matching step-up transformer but at a cost.

2.6 Static Var Generator Converter (SVC)

This configuration is shown in Fig 7 [2]. It has remarkable properties such as a sinusoidal current being drawn at unity power factor. The three phase PWM inverter is phased locked to the supply and the angle and voltage magnitude are controlled so as that the converter supplies only reactive energy whilst maintaining the supply at unity power factor. The converter has lots of merit, however the voltage utilisation appears to be low. It would also require a step up transformer if standard motors were to be used. Nevertheless it is certainly an interesting topology and may well find applications in Flexible Transmission Systems (FACTS).

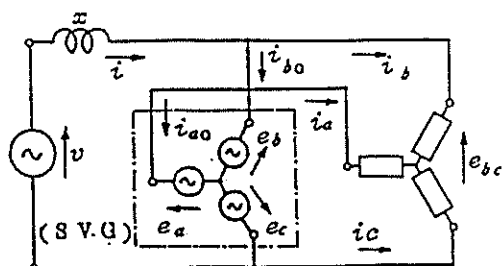


Figure 7: Single phase to three phase using SVC

3 Proposed Converter

[illegible]

Figure 8: Circuit to produce balanced three phase with reduced size inverter locked at 90 degree to supply.

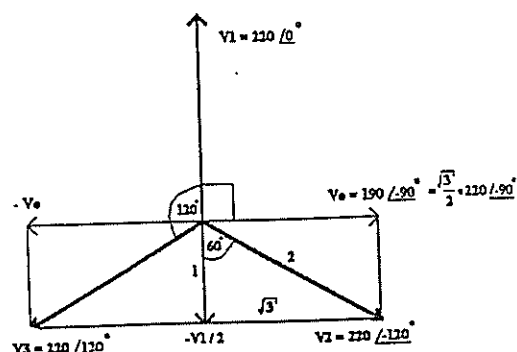


Figure 9: Phasor diagram for above circuit

therefore $V_{O10} = 3/2 V_{I1}$. This is half the total output rating. The method hence not only simplifies the requirement of the inverter to a single phase one but also only needs to be half the size. Pspice simulation with experimental results showed a supply power factor of .72 with a .8 power factor load and filter inductance L_s of 1 mH and capacitor C_s of 10,000 micro Farads. The THD is in the order of 20% but is expected to be drastically improve with a three phase rectifier fed from the output of the converter instead of the single phase rectifier shown in Fig 8. This is the topic of further work.

4 Conclusion

The proposed converter has distinct advantages. These are that the voltage utilization is high and enables a standard 3 phase motors to be used. The semiconductor inverter is a single phase and is half the size of the output rating. The single phase supply power factor and current distortion is reasonable and can be much improve if a three phase rectifier fed from the output is used this will be the subject of future work.

5 Acknowledgements

Acknowledgement is given to UCT students M.A. Khan and I.Clegg for experimental work and circuit implementation and also to MLT Drives for loan of equipment and assistance.

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Design of Robust SISO Excitation Controllers Based on H_∞ Control Theory

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Abstract: The problem of weakly damped power oscillations in longitudinal power systems is well known and the design of Conventional Power System Stabilizers (CPSS) in the form of lead-lag circuits has been actively researched. These CPSS are however, only valid for a small region about the operating point and they possess very weak robustness and sensitivity properties. This paper presents a linear design method based on H_∞ control theory to design robust excitation controllers. The controllers are designed with sensitivity and robustness margins explicitly specified in the selection of appropriate weighting functions. The paper makes use of SISO excitation control to damp power oscillations. The H_∞ -based controller is shown to possess superior robustness properties to that of the CPSS.

1. Introduction

Conventional generator excitation control systems consist of Automatic Voltage Regulators (AVR) for transient voltage regulation and Power System Stabilizers (PSS) for damping of Mechanical Mode Oscillations (MMO). The design of these lead-lag excitation controllers is based on classical linear systems theory and is therefore valid for only a small region about the operating point. In addition, conventional excitation control design uses single variable control theory for controlling a system that is multivariable in nature.

Recently, there have been reports demonstrating the poor performance of conventional PSS under both transient and time varying system operating conditions. The PSS have been shown to cause a large overshoot in the terminal voltage during removal of the fault. In addition, under light loading conditions PSS can provide inadequate damping of the generator thus resulting in unacceptable oscillations under small disturbances [1-2].

One way of addressing these problems is to make use of non-linear control theory. Several adaptive controllers have been proposed as replacement to conventional excitation control [1-3]. However, these techniques still suffer from serious drawbacks such as the difficulty of ensuring stability of the controller.

An alternative strategy is to make use of linear systems theory that minimizes the performance of sensitivity to changes in the model of the plant. This paper makes use of modern linear control theory in the form of H_∞ optimal control to design robust excitation controllers. The design is limited to the voltage loop since control energy in the power loop must be minimized.

The controllers are designed on a nominal SMIB system and tested under a range of loading conditions. Time domain simulation results are presented. The performance of the H_∞ controller is compared to that of the CPSS. The CPSS used in the comparison was designed using a pole-placement technique [3]. The simulation results indicate that the system with the H_∞ controller has satisfactory closed loop response under a range of operating conditions. The H_∞ controller is shown to possess superior robustness properties compared to that of the CPSS.

2. H_∞ Design for Excitation Control

2.1 Analysis of Network Under Investigation

Figure 1 shows a diagrammatic representation of the system under investigation.

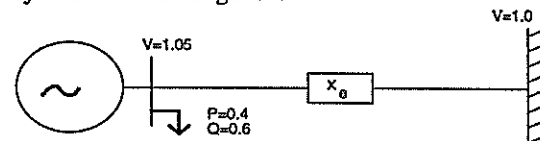


Figure 1: System Under Investigation

The system consists of a single machine connected to an infinite bus by means of a transmission line. The nominal operating point is shown in Figure 1. The generator control system consists of the AVR for voltage control and the speed governor for speed control. No PSS is modeled. The generator is modeled as a sixth order model, the excitation system as a second order model and the governor as a third order model. The load is modeled as a constant impedance load and the network equations are algebraic.

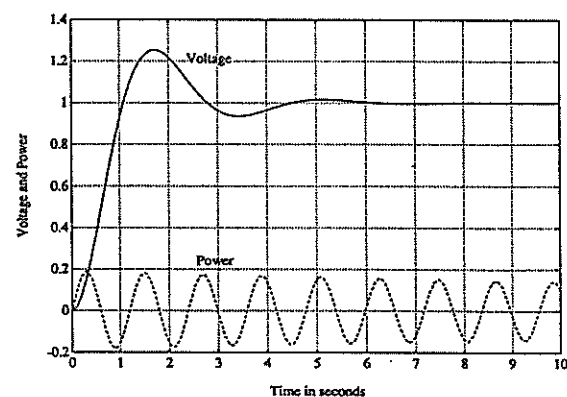


Figure 2: Open Loop Step Response in Voltage and Power for a Step in V_{ref}

Figure 2 shows the open loop step response of the system under study. It can be seen that the voltage loop has unsatisfactory transient performance due to the large overshoot and long settling time. In addition, there exists a large interaction between the voltage and power loop. The power oscillations that result have very weak damping. This can be seen from the selected poles shown in Table 1.

Poles	Zeros
$-0.336 \pm j 5.259$	$-0.362 \pm j 5.566$
$-0.770 \pm j 1.378$	-1.036
-1.080	-2.000
-8.217	-16.00

Table 1: Poles and Zeros of the SMIB System

The complex conjugate poles in row 1 are the mechanical mode eigenvalues obtained from the swing equation. The damping factor for these eigenvalues is small and the system thus requires supplementary control. Since this paper is concerned with excitation control the zeros of the voltage loop are of interest. As can be seen from Table 1, there exist a complex conjugate pair of zeros close to the mechanical mode poles. This pair of zeros will be called the Mechanical Mode Excitation Zeros (MMEZ).

The aim of the H_∞ method used in this paper is to design a controller for which the closed loop system has satisfactory response in the voltage loop and for which the power oscillations are well damped.

2.2 Design Specifications for SMIB

The required response of the voltage loop of the SMIB system needs to be specified. This can be summarized as follows:

- The closed loop time constant must be equal to or slightly less than that of the open loop.
- The bandwidth of the closed loop system should be limited so as to avoid excitation of high frequency noise signals.
- The steady state error must be less than 1%.

The voltage loop specifications can be translated into frequency domain constraints in the form of three weighting functions W_1 , W_2 and W_3 [7]. Table 2 shows the parameters of the weighting functions selected for the voltage loop supplementary control.

Sensitivity (W_1)	$(s + 100)/(100s + 1)$
Input Energy (W_2)	0
Robustness (W_3)	$s^2/1000$

Table 2: Parameters of Weighting Functions

2.3 Optimal H_∞ Solution

The H_∞ algorithm is used to solve for the controller that satisfies the constraints of the weighting functions. The algorithms in PC-Matlab™ were used for this purpose. The optimal solution is found by one-parameter (γ) iteration and was obtained by the method of bisection. The results obtained for the optimal γ search for voltage loop control is shown in Appendix A. The closed loop voltage response is shown in Figure B1. A comparison of the roots given in Table 1, and Table A1 demonstrates some features of the H_∞ synthesis methodology. These are:

- The poles of the H_∞ controller approximately equal the zeros of the open loop plant thus canceling the effect of the open loop zeros.
- The zeros of the H_∞ controller approximately equal the poles of the open loop thus canceling the open loop poles.

From the above, it can be seen that the H_∞ method achieves its design objectives by affecting the observability of the open loop poles. However, it does not affect the damping factor of these poles. Thus, if the system operating point changes, the poles and zeros will move apart from each other. The weakly damped poles will therefore begin to dominate the performance of the closed loop under such changing operating conditions.

In this paper, we wish to demonstrate the use of excitation control for damping of power oscillations. It may appear therefore that the H_∞ method of pole-zero cancellation explored above does not offer a solution to damping of power oscillations since only the observability of the voltage loop poles are affected. The poles of the power loop will remain unaffected. In the next section however, we describe a technique to improve the damping of the power loop while still making use of SISO excitation control.

2.4 Improving Power Loop Damping Via Voltage Loop Control (Sub Optimal H_∞)

The damping of power oscillation will be restricted to control via the voltage loop. This is because fast control through the power loop is mechanically stressful on the valves and therefore undesirable.

Figure 3 shows the method that was used to improve the power loop damping via voltage loop control. In Figure 3a the pole-zero plot of the SMIB voltage loop is shown (only selected poles and zeros are shown). By means of a bilinear transformation the $j\omega$ -axis is shifted passed the axis of the mechanical mode poles. This creates a fictitious unstable plant $g'(s)$. The standard H_∞ algorithm is then applied to $g'(s)$.

The zeros of the H_∞ controller $k(s)$ that correspond to the unstable poles of $g'(s)$ are placed at the reflected positions about the $j\omega$ -axis (Figure 3c). The closed loop pole-zero plot (Figure 3d) shows the shift in the unstable eigenvalue past the $j\omega$ axis of $g'(s)$.

Thus, $g'(s)$ is stabilized and the damping of the poles of $g(s)$ is improved. The resulting controller is optimal for $g'(s)$ but sub optimal for $g(s)$. Note that the robust stability margin of $g(s)$ is improved by the sub optimal controller since the weakly damped poles are shifted away from the $j\omega$ axis.

By applying this simple technique, the damping of the mechanical mode oscillation can be enhanced via SISO excitation control.

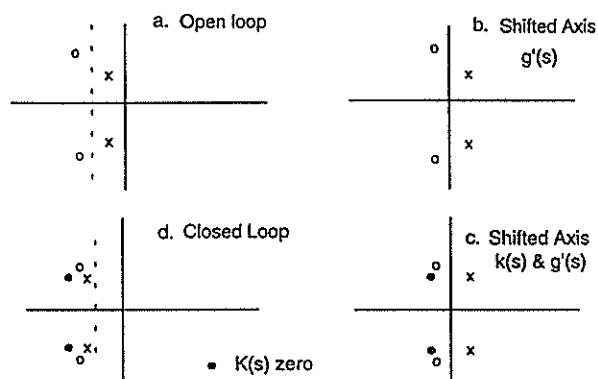


Figure 3: Shifting the $j\omega$ axis to improve damping

From Figure 3 it can be observed that the amount of damping that can be achieved by shifting the axis is limited by the pole-zero pattern of the open loop voltage loop. If the $j\omega$ axis is shifted past the axis of the MMEZ then $g'(s)$ will become an unstable non-minimum phase plant (will possess poles and zeros in the right half plane). For such plants, it is only possible to find a stable stabilizing controller if the open loop plant is Strongly Stabilizable (SS), or equivalently, if the plant satisfies the Parity Interlacing Principle (PIP) [10].

For unstable non-minimum phase plants the H_∞ algorithm often provides unstable controller's $k(s)$. However, it has been shown that the selection of high order weights together with a sub optimal γ search ensures a stable $k(s)$ [10].

Figure 4 shows the flowchart of the stages in the design procedure for damping of power oscillations via voltage loop control. The flowchart incorporates all the necessary adjustments to the H_∞ algorithm. The resulting H_∞ controller is a stable sub optimal solution.

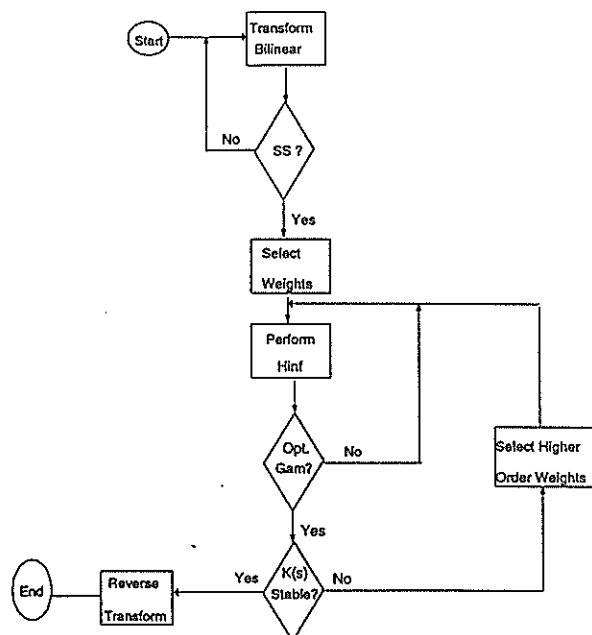


Figure 4: Flowchart showing the procedure for sub optimal H_∞ design

Appendix B shows the time domain step response for the closed loop system. Both the H_∞ -based controller and the CPSS closed loop systems are shown. The operating point is changed and the same comparison of the two controllers shown. In all the cases the H_∞ controller has better performance than the CPSS. The most significant feature is that the H_∞ controller possesses superior robustness properties in the face of changes in loading conditions.

3. Conclusion

Robust H_∞ controllers were designed for a SMIB system in order to damp mechanical mode oscillations in the power loop. In order to improve the damping and robust stability margin of the closed loop it is necessary to make use of a bilinear transformation. This in turn necessitates the use of high order weights to ensure the existence of a stable controller. The resulting controller is a sub optimal H_∞ controller. The sub optimal controller is always stable and has an increased robust stability margin as compared to the standard optimal H_∞ solution. The sub optimal H_∞ controller response was shown to be superior to that of the CPSS under varying load conditions.

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5. Appendix

A. Pole/Zero Pattern of H_∞ Controller

Controller Poles	Controller Zeros
$-0.362 \pm j5.566$	$-0.336 \pm j5.259$
-1.030	-8.217
-2.000	$-0.77 \pm j1.378$
-7.798	-1.080

Table A1: Poles and Zeros of Controller k(s)

B. Simulation Results

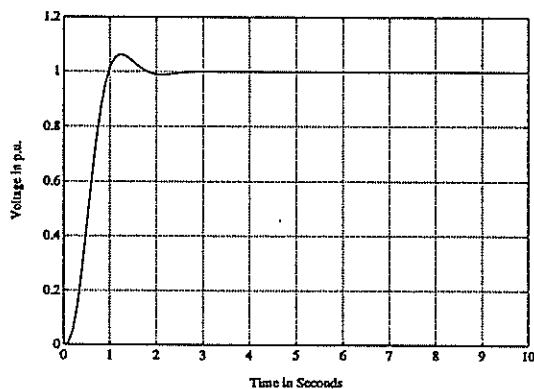


Figure B1: Closed Loop Voltage Response

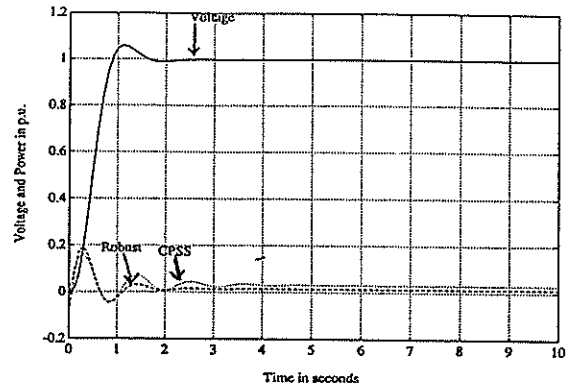


Figure B2: Step Response (Nominal Load)

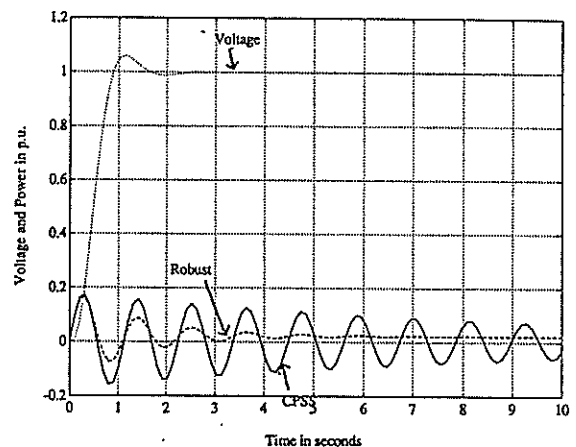


Figure B3: Step Response (20% Increased Load).

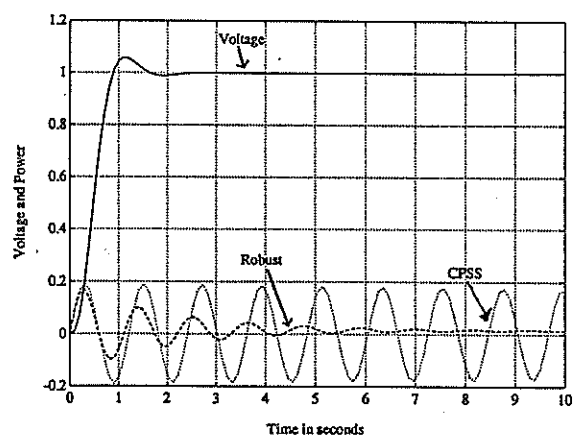


Figure B4: Step Response (20% Reduction in Load).

THE DESIGN OF FEEDBACK CONTROLLERS BASED ON QUALITATIVE FEEDBACK THEORY FOR SWITCH MODE POWER SUPPLIES

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Abstract. A nonlinear discrete-time model of the Buck-Boost converter is linearised to obtain a linear discrete-time model. If the duty cycle of the Buck-Boost converter is changed, then a different linearised model is obtained. These linearised models can then be used to quantify the variation in the dynamics of the Buck-Boost converter with changing duty cycle. This analysis is an important step in the design of controllers based on Qualitative Feedback Theory (QFT). These controllers can be designed to give the required closed-loop operation and stability for all operating conditions. Even though a simple circuit is used this method is suitable for more complicated switching circuits. It is also shown that parasitic elements are adequately accounted for.

Keywords. DC-dc Converter, Qualitative Feedback Theory, Parameter Variation, Linear Discrete-time model

INTRODUCTION

Dc-dc converters are widely used in regulated switch-mode dc power supplies and in dc-motor drive applications. In a dc-to-dc converter, for a given input voltage, the average output voltage is controlled by controlling the switch on and off durations (t_{on} and t_{off}). This paper assumes Pulse Width Modulation (PWM) switching, with a constant switching frequency (hence, a constant switching time period $T_s = t_{on} + t_{off}$) and a varying duty ratio D , which is defined as the ratio of the on-duration to the switching time period ($D = t_{on}/T_s$). The duty ratio D is varied to control the average output voltage.

Regulated switch-mode dc power supplies use negative feedback to maintain the average output voltage constant, in the face of disturbances, model uncertainty and parameter variation. Figure 1 shows a typical control system configuration. Measurements that reveal the circuits present behaviour are used to choose a suitable control action aimed at restoring nominal operation rapidly and safely. Deviations from the nominal operating condition are called the dynamic behaviour of the circuit. Controller design and stability analysis requires the appropriate dynamic models. These dynamic models describe how control action and disturbances will affect the future behaviour of the circuit.

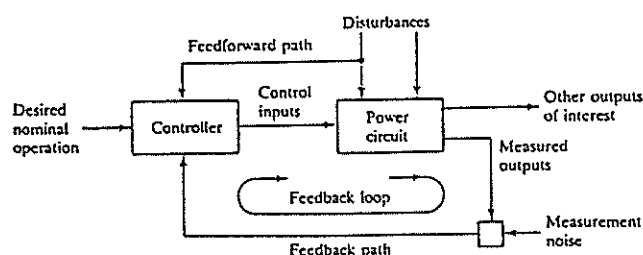


Figure 1 : A typical control system configuration (after Kassakian et al(6))

Dynamic models of power electronic circuits are typically nonlinear. This makes the design of controllers and stability analysis very difficult. Linear time-invariant models can, however, be obtained by linearisation of the nonlinear model about a nominal operating point. These models can be used to approximately describe small deviations or perturbations from nominal operation. This then allows for the use of classical control theory in the design of controllers and for the evaluation of stability.

In power electronic circuits, however, the dynamic models involve periodically varying quantities even at steady state. One way to get a constant steady state is to use a discrete-time model that uses circuit variables sampled once per cycle. The averaged models discussed by Middlebrook et al (1) also yield constant steady state, but are not as accurate. The nonlinear discrete model can be linearised about a nominal operating point to obtain a linear time-invariant discrete-time model. This model is dependant on the operating point and is only valid for small perturbations from the operating point.

Controllers developed from linearised models cannot be guaranteed to provide satisfactory operation for large deviations from nominal operation, and stability can be a problem. Also classical control design cannot account for model changes with changes in operating point (for instance changes in duty cycle).

Qualitative Feedback Theory (QFT), however, uses negative feedback to achieve the desired system performance despite plant uncertainty, model changes and plant disturbances. This is done by quantifying the model variation or uncertainty and the plant disturbances. A controller can then be designed to give the required operation and stability for *all operating conditions*.

This paper quantifies the model variation due to changes in duty cycle (or operating point) using the discrete-time model of the Buck-Boost dc-to-dc converter. Even though a simple circuit is used this method is suitable for more complicated switching circuits.

A brief outline of the general discrete-time model is given in Section 2. Section 3 discusses QFT and a few associated problems. Section 4 gives results obtained for the Buck-Boost converter.

THE DISCRETE MODEL

A general discrete-time model of arbitrary power electronic circuits has been proposed by Verghese et al (2) and von Lutz et al (3). This leads to linearised small-signal models that describe perturbations about a nominal cyclic steady state. Switches in power electronic circuits change mode at discrete points in time which are either state dependant, as for the diode, or specified, as for the power transistor. The former are referred to as 'state dependant events' while the latter are 'time dependant events'. The times at which the events occur will affect the characteristics of the system. The general nonlinear discrete-time model of a power electronic circuit can be described by a nonlinear function

$$x(k+1) = F(x(k), T(k), k) \quad (1)$$

$$k = 0, 1, 2, \dots$$

where x is the vector of states of the circuit and T is the vector of time dependant events. The time dependant events can be controlled by the user and can therefore be considered as inputs to the system. The exact solution of (1) is seldom possible. The function $F(\dots)$ can, however, be linearised about a steady state operating point x^* , T^* , giving a perturbation model of the form

$$\Delta x(k+1) = A \Delta x(k) + B \Delta T(k) \quad (2)$$

where Δx and ΔT are the perturbation of x and T about x^* and T^* respectively. The matrix A represents the sensitivity of the state variable $x(k+1)$ on the state variable $x(k)$ and B represents the sensitivity of $x(k+1)$ on $T(k)$. Mathematically these matrices are expressed as follows:

$$A = \left. \frac{dF(x(k), T(k), k)}{dx(k)} \right|_{x^*, T^*} \quad (3)$$

$$B = \left. \frac{dF(x(k), T(k), k)}{dT(k)} \right|_{x^*, T^*}$$

This linearisation procedure has been incorporated into a simulation package called CASED (Computer Aided Simulation of Electric Drives) developed by McCulloch (4). CASED simulates the circuit operation to obtain the periodic steady state conditions, and then calculates the A and B matrices for the linearised discrete-time model.

The following should be noted with respect to the discrete model.

- This is a general model that can be used for any power electronic switched circuit.
- Equation (2) describes a perturbation model. This implies that some nominal switching strategy must have been used to obtain steady state. This technique will not generate a switching strategy, it can only optimise the already existing one.
- The model describes the behaviour of the system at discrete points in time. The behaviour between sampling instants is therefore not known. The sampling period of the model is normally taken to be equal to a cycle of the converter.

QUALITATIVE FEEDBACK THEORY

QFT is a two-degree-of-freedom design technique, using negative feedback and a prefilter $F(s)$ to achieve the desired system performance. See Figure 2. This technique requires that the model variation due to parameter variation, changes in operating point and model uncertainty be quantified over the bandwidth of interest. The loop transmission, $L(s) = G(s)P(s)$, is made as large as possible over the desired bandwidth so that the sensitivity to parameter variation is as low as possible and the disturbance rejection as high as possible. $L(s)$ must then roll-off as fast as stability constraints will allow, to minimize the effect of noise. This is achieved via the design of $G(s)$, the first degree-of-freedom. The resulting closed-loop transfer function will be close to unity over the design bandwidth. The second degree-of-freedom, $F(s)$, is designed to adjust the overall transfer function to that required.

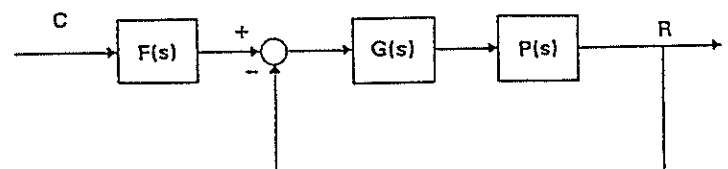


Figure 2: A typical two-degree-of-freedom system

The price paid for this reduction in sensitivity is an increased loop bandwidth. If the plant has a large amount of uncertainty then a large bandwidth is required to cope with the accurate control of the system. The

problem then is that of sensor noise. Noise amplification can be reduced by using multiple feedback loops or adaptive control.

Adaptive control assumes large slow parameter variations. Observers are used to estimate the plant dynamics and to adjust the gain suitably. In dc/dc converters we have the usual slow parameter variations of the circuit components. A change in duty cycle, however, to compensate for a change in output voltage, rapidly changes the circuit model. Adaptive control cannot be used since the rate of parameter variation is too large when the duty cycle changes.

Multiple feedback loops, measure internal variables, resulting in individual loops having lower bandwidths than the single loop case. This requires more sensors, but the 'burden' of feedback is distributed between the loops. In the Buck-Boost converter multiple loop feedback can be used to reduce noise amplification if necessary, by feedback of the inductor current. Discrete-time models can be generated, using CASED, to quantify the model variation of inductor current with changes in duty cycle. A QFT controller can then be designed for what is usually called Current Mode Control.

The design may still be unattainable because of the plant parasitics or if the plant is nonminimum phase.

APPLICATION TO A BUCK-BOOST CONVERTER

The Buck-Boost converter chosen for analysis is shown in Figure 3. The component values are given in Appendix A. The linearised models obtained from CASED were used to analyse the dynamics of the converter and to quantify the model variation for duty cycles between 0 and 1.

Figure 4 shows the pole migration in the discrete-time z complex planes for varying duty cycle at a constant switching frequency of 6kHz. As the duty cycle increases the converter dynamics change from being overdamped to underdamped.

Figures 5 and 6 show the bode plots (Gain and Phase respectively) of the transfer functions $\Delta V_{C_{ap}}(k)/\Delta T_2(k)$ for duty cycles between 0 and 1. $T_1(k)$ and $T_2(k)$ are time dependant events, which can be controlled and are therefore inputs to the system. See Figure 7. For a constant switching frequency T_1 is constant, while T_2 which represents the duty cycle is controlled. The state variable $V_{C_{ap}}$ is the output voltage.

The upper and lower bounds of the magnitude plots represent the open-loop model uncertainty for all operating conditions. Parasitic elements have a significant impact on the voltage conversion ratio as shown by Mohan et al(5). Figure 8 shows the effect of these parasitic elements. The phase plots confirm this effect, with a 180° phase shift for high duty ratios. This

may lead to instability in the feedback loop if not accounted for.

If multiple loop feedback is required bode plots similar to those shown in Figures 5 and 6 can be generated, to quantify the model variation of inductor current with changes in operating point.

CONCLUSIONS

The model variation due to changes in duty cycle has been quantified, for the Buck-Boost dc-to-dc converter, using the exact linearised discrete-time models. Emerging from this analysis is a novel method of designing feedback controllers using QFT. These controllers can be designed to give the required closed-loop operation and stability for all operation conditions. To reduce noise amplification feedback of internal variables can be used resulting in multiple loop feedback.

Even though a simple circuit is used this method is suitable for more complicated switching power electronic circuits.

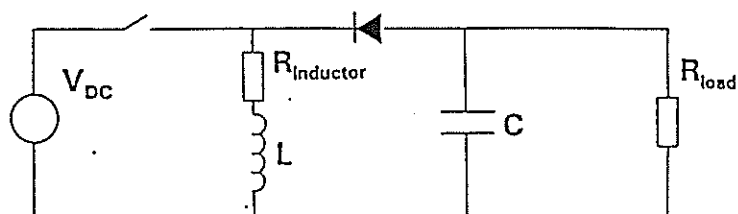


Figure 3 : Buck Boost DC-to-DC converter

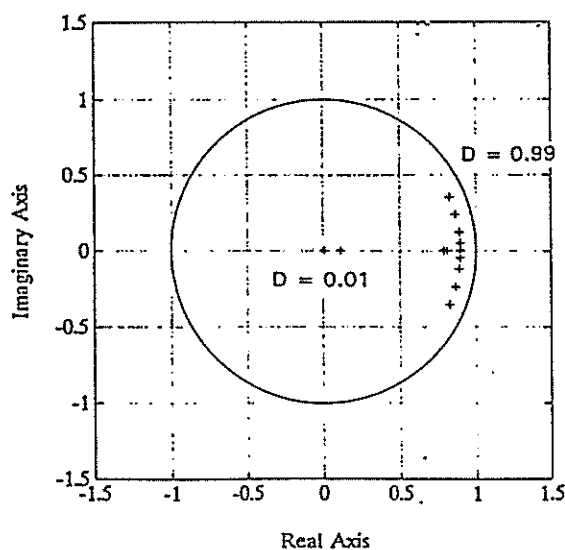


Figure 4 : Locus of poles in the Z -plane for varying duty cycle ($f_s = 6\text{kHz}$)

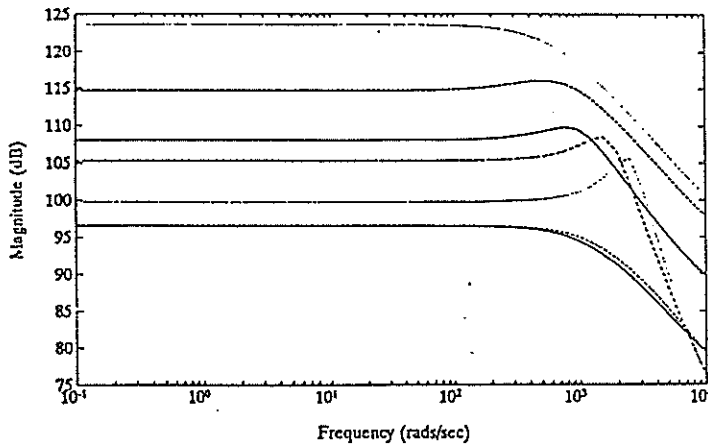


Figure 5 : Magnitude plot of small-signal model $(\Delta V_{Cap}(k)/\Delta T_2(k))$

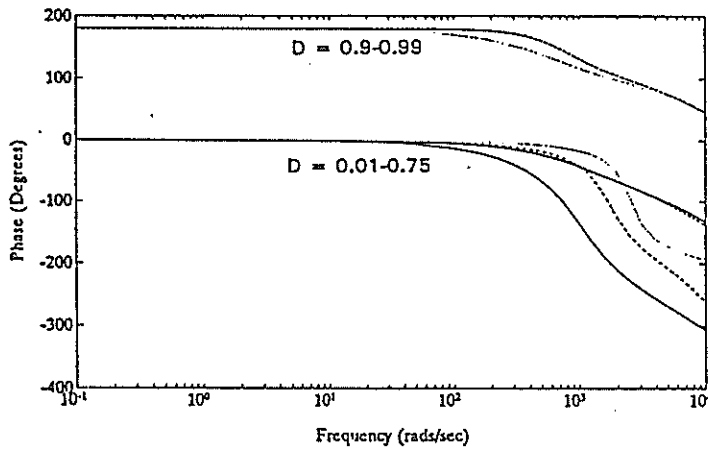


Figure 6 : Phase plot of small-signal model $(\Delta V_{Cap}(k)/\Delta T_2(k))$

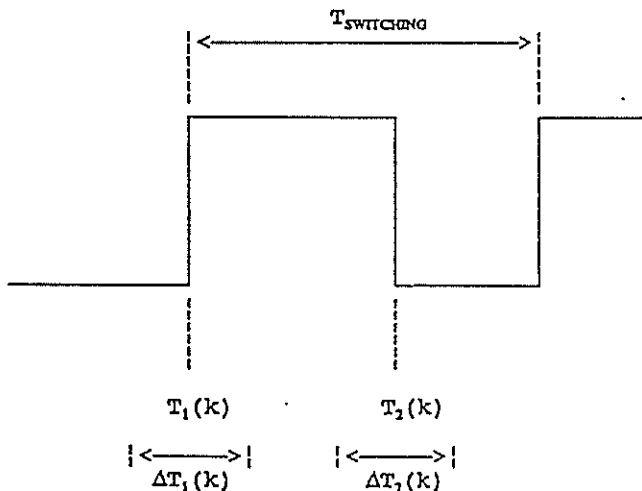


Figure 7 : Switch control signal

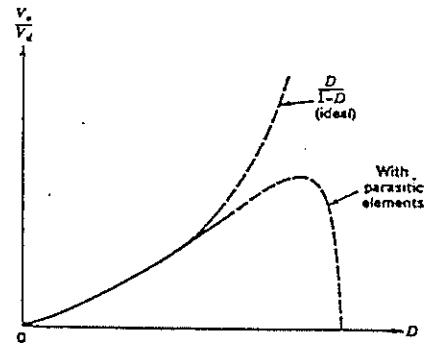


Figure 8 : Effect of parasitic elements on the voltage conversion ratio in a Buck-Boost converter (after(5))

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APPENDIX A

The Buck-Boost converter component values are as follows:

$$\begin{aligned}
 R_{\text{inductor}} &= 0.652 \, \Omega & C &= 80 \, \mu\text{F} \\
 V_{\text{DC}} &= 10 \, \text{Volts} & L &= 1.2 \, \text{mH} \\
 R_{\text{diode on}} &= R_{\text{transistor on}} = 0.158 \, \Omega \\
 R_{\text{diode off}} &= R_{\text{transistor off}} = 10 \, \text{M}\Omega \\
 R_{\text{load}} &= 20 \, \Omega
 \end{aligned}$$

THE TRANSFER FUNCTION AND TRANSFORMER IMPULSE TESTING

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Abstract: During the last decade HV (High Voltage) laboratories have been switching over to digital monitoring equipment. This provides the opportunity for the use of signal analysis techniques. One of the techniques recently being examined is the transfer function method. This method uses the sampled test data of the input impulse test voltage and resulting output current from the 100% and 50% BIL tests. An exiting feature of this method is the possibility to determine the locality and type of fault occurring in the transformer.

1. Introduction

The potential benefit of digital recording in the field of HV is discussed by R Malewski and B Poulin [1]. They compared the resolution of the conventional analog impulse oscillogram to that of a commercial digitizer in the bandwidth corresponding to the frequency spectrum of the recorded transients and found that the digital device can be applied for monitoring HV impulse tests instead of using a conventional oscilloscope.

The method of system characterization by the analysis of its frequency behaviour is called Frequency Response Analysis (FRA). This low voltage frequency response analysis could only be used for the study of linear effects like mechanical changes in the interior of the transformer because these affects modify the internal distribution of capacitance and inductance and thus the frequency response. This method is not usable for voltage related phenoma such as corona and partial discharge.[2]

Voltage and current signals, acquired at a reduced test level (normally 50%) of the maximum test voltage, also called Basic Insulation Level (BIL) and at the full test level are compared to determine whether the transformer is healthy. This is done under the assumption that the transformer is a linear test object. Any discrepancies showing on the full test data can thus be said to represent a fault condition. This method required the test voltage to be the same in shape for every test and certain faults in the impulse generator can thus condemn a healthy transformer. As the transfer function of each transformer is unique, the shape of the input voltage form is not important with regard to the transfer function.

2. Theory

Digital hardware and the fast Fourier Transform (FFT) are not sufficient to ensure successful signal processing. The user must also understand the measurement process itself.[3] The measurements must be unambiguous and quantitative, which is only possible if it yield numerical values that can attach to a give aspect of the system with only one interpretation. The correlation and transfer functions are thus briefly discussed.

Consider a generalized measurement situation as shown in Figure 1 where there are two accessible measuring points X and Y. X and Y are related by some linear transfer function H while Y is perturbed by some uncorrelated noise source N . The two basic measurements involve the transmission path H and the degree to which the uncorrelated noise N affects the output Y . The first is a measurement relating Y and X , the second the degree of causality between X and Y .

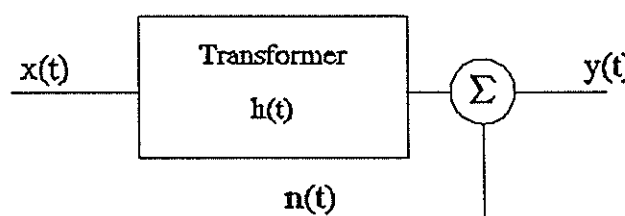


Figure 1. Model to simulate determination of the transfer function using correlation techniques

Consider the two functions that can be used to measure the relationship between two signals. The first is an estimation of the correlation between X and Y over a period of time T given by equation 1.

$$R_{yx}(\tau) = \int_0^T y(t) x(t+\tau) dt \quad (1)$$

Where R_{yx} is simply the averaged product of y lagged with respect to x . When the value of the cross correlation is high for some value of the lag τ it can be said that x and y are similar. The cross

$$\rho_{yx}(\tau) = \frac{R_{yx}(\tau)}{\sqrt{R_{xx}(0) R_{yy}(0)}} \quad |\rho_{yx}(\tau)| \leq 1 \quad (2)$$

correlation is normalized by dividing by the geometric mean of the input as well as the output mean-squared value, given by equation 2. The cross power spectrum, which is the Fourier transform of the cross-correlation function, is the key element in this approach, eq 3

$$G_{yx} = \mathcal{F}[R_{yx}] \quad (3)$$

The input power spectrum and output power spectrum are thus formed. (Eq 4 and 5)

$$G_{xx} = \mathcal{F}[R_{xx}] \quad (4)$$

$$G_{yy} = \mathcal{F}[R_{yy}] \quad (5)$$

The output power spectrum could be split into two components. One is due to the input while the other results from the uncorrelated noise source. The cross power spectrum could thus be measured

$$G_{yx} = H G_{xx} + G_{nx} \quad (6)$$

The first term represents a measurement of the transmission path, while the second term is related to the degree of noise contamination. Two types of measurement errors are found, namely the deterministic and variance error. Both the deterministic and variance error can be nearly eliminated if the transfer function is calculated by dividing the input power spectrum by the cross power spectrum. The transfer function is thus

$$H = \frac{\overline{G_{yx}}}{\overline{G_{xx}}} \quad (7)$$

The only precaution is that G_{xx} and G_{yx} are from simultaneously sampled records. If not, the random variation would not be the same in each quantity and the cancellation would not occur.

3. Experimental results

A 33kV transformer winding was subjected to impulse testing. The test setup is shown in figure 2. In the figure the symbols means the following: C_s is the serial capacitor while R_{cs} is the losses associated with it. L_s is the induction of $1/N$ section of the winding. R_s is the corresponding conductor resistance. C_p is the parallel capacitor while R_{cp} represents the associated parallel

capacitor losses V is the voltage divider and R_{shunt} is the coaxial shunt for current measurements. $S1$ represents the defaults.

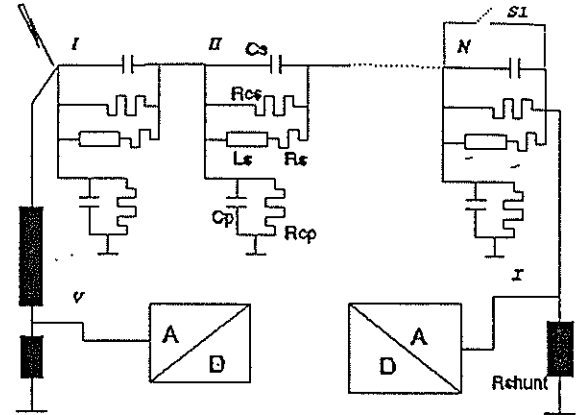


Fig. 2 Transfer function test setup

The winding was subjected to a chopped impulse as shown in figure 3. The current resulting from this impulse is shown in figure 4. The windings was systematical short

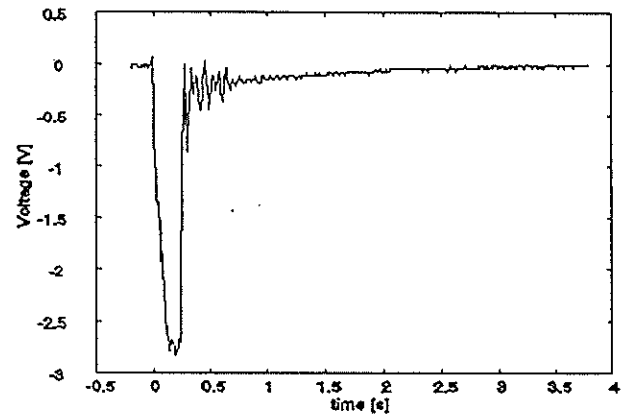


Fig. 3 Test voltage

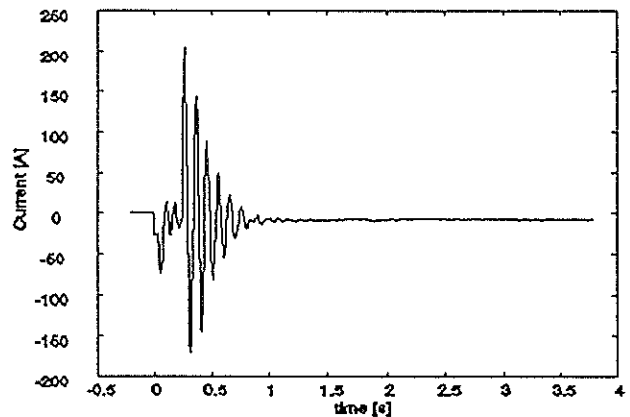


Fig. 4 Measured current

circuit to simulate an interturn short circuited. The procedure outlined above was used to determine the transfer functions. Due to the voltage distribution over the

winding the difference between the obtained transfer functions with phase are shown relatively to the healthy transformer. The transfer functions are shown in figure 5a to figure 7a. The phase is shown from figure 5b to figure 7b. In each of these figures 5 to 7 the solid line is the healthy function while the perturbed line is that of the simulated fault. To ensure the longevity of the winding the test was performed at a lower voltage than rated.

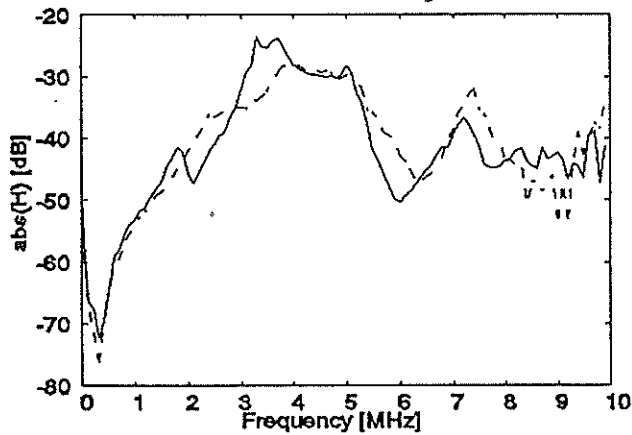


Fig. 5a Transfer function : Winding 1 and 2 short circuited

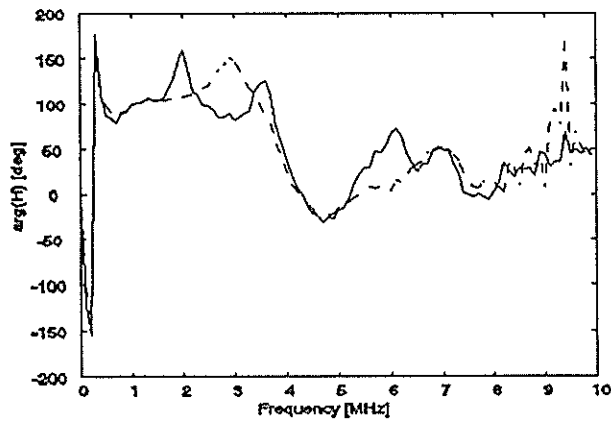


Fig. 5b Phase plot: Winding 1 and 2 short circuited

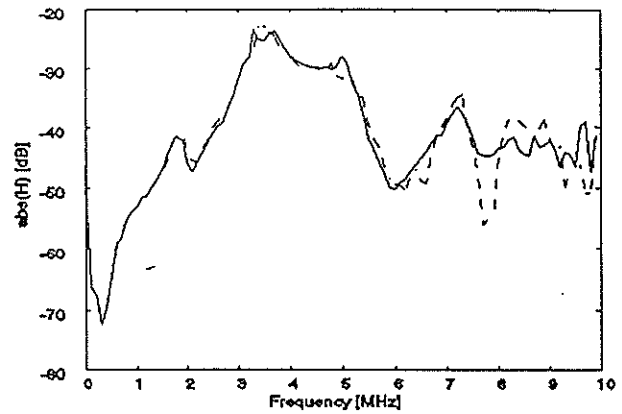


Fig. 6a Transfer function: Winding 19 and 20 short circuited

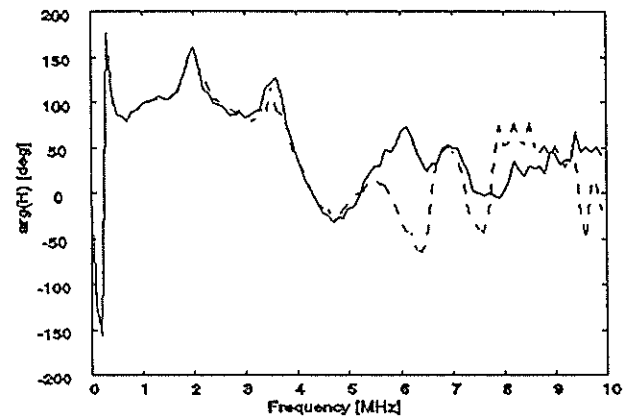


Fig. 6b Phase plot: Winding 19 and 20 short circuited

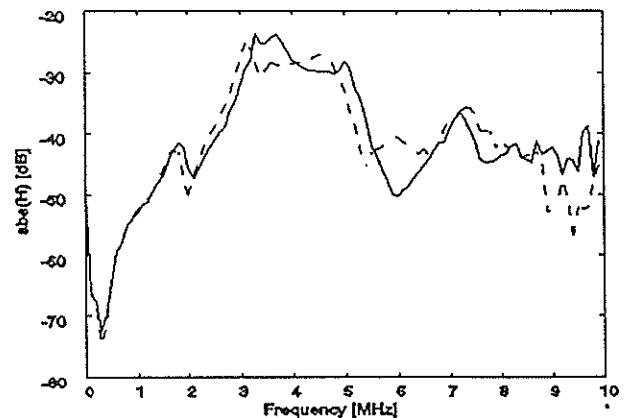


Fig. 7a Transfer function: Winding 39 and 40 short circuited

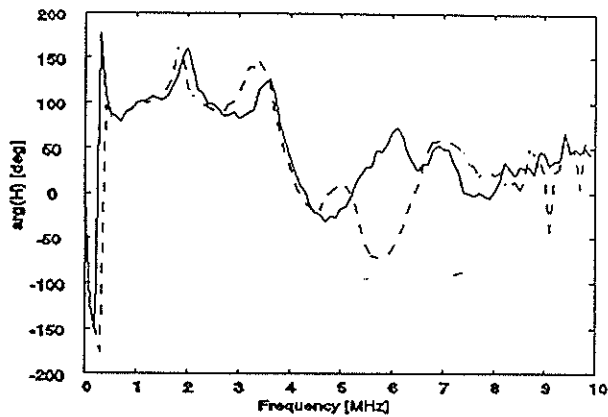


Fig. 7b Phase plot: Winding 39 and 40 short circuited

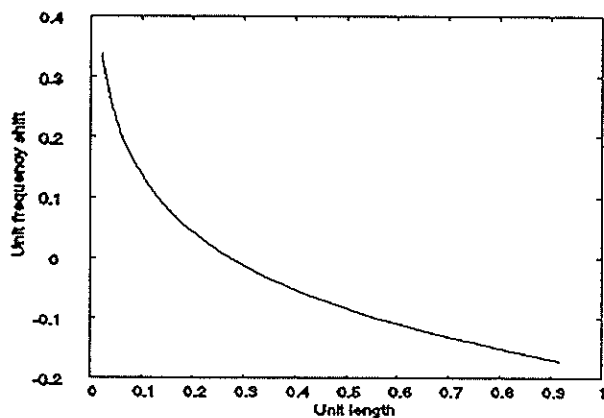


Fig. 8 Frequency shift as function of fault position

4. Discussion

From the transfer function plots the difference between the simulated faults and the healthy transformer could clearly be distinguished. An interesting occurrence is the shifting of the poles from far right to left of the healthy function with regards to the position of the fault (fig. 8). This tendency is closely copied in the phase plot.

This difference also reflects on the magnitude of the transfer function. For faults higher up in the winding more discrepancies exist at the low frequency than for faults lower in the winding. The opposite is true for the high frequency region where the discrepancy is more for faults in the lower region.

5. References

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HARMONIC LOAD FLOW IN MEDIUM VOLTAGE NETWORK

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ABSTRACT

The harmonic load flow has been analyzed in a part of Port Elizabeth network. The harmonic distortion of the voltage and current waveform is represented with total harmonic distortion THD and RMS coefficients. The phase angle of the harmonic source is varied from -90 to $+90$ degrees and this issue is taken into account during the analyses and the prediction of harmonic distortions. The consequence of load current to short circuit currents ratio to THD is observed. Compensation of power factor and harmonic distortion should be done for the group of consumers due to better technical and financial effects. Taking harmonic power flow through the network into account, the devices are usually installed at the main distribution substation.

1. INTRODUCTION

Quality of electrical supply affects all industrial consumers as well as each commercial and domestic consumer. Ideal network would consist of steady frequency and unspoiled sinusoidal shape of voltages and currents but many factors caused by the source, transmission lines and the load itself distort these theoretical values. These deviations can lead to maloperation of computerised automatic or other electronic equipment. It also induces unexpected currents and overvoltages at telecommunication cables and equipment. For this reason the authorities, as suppliers, should react on interference of one consumer with another due to quality of supply.

There is a need for a simple way to determine a distortion of an existing system. Growing number of theories analyze harmonic distortion. Some simplified methods with probabilistic approach are used for a quick check. Harmonic theory approach to determine the distortion and quality of supply simplifies the calculation even though hides the reality.

The characteristic of a nonlinear load changes the sinusoidal nature of the AC power current and therefore produces a voltage drop resulting in the flow of harmonic currents.

Few aspects must be taken into consideration while studying problem of harmonic flow through the network:

- model of a system, node location
- load representation and seasonal variation
- presence of harmonic filters within consumers

The most significant sources of harmonic distortions are static power converters, arc discharge devices and saturated magnetic devices. Static power converters include large silicon controlled rectifiers, power converters and diode rectifiers. Characteristic harmonics for the converters depend on switching combination and number of diodes as :

$$k = q \pm 1$$

k - harmonic order

q - converter pulse (depends on number of diodes)

Since new loads are being connected all the time, the distributor should control the quality of supply and set or adopt the limits of distortion. That would prevent the consumer, potential source of harmonic distortion, to be connected if out of limits.

Measurements are essential to support the theoretical approach and justify the results obtained by means of computers.

Instruments are placed at all points of interest from the harmonic load flow point of view particularly at the points of common supply and main distribution busbars.

Harmonic flow in a system and value of harmonic distortion depend on characteristics of all elements such as capacitors, loads, switching status change, size and impedance of transformers.

Checking of harmonics is a continuous process since the system changes.

2. ANALYSES

The system under consideration is 132/66/22kV Port Elizabeth network. Main supply to the network is a double line from ESCOM's Grassridge 220/132 kV substation to Chatty as it is indicated in Figure 1. Chatty is the main distributive 132 kV substation with parallel lines running to the industrial area. The 132/66 kV stepdown switchgear at Swartkops provides three feeder-transformers to the main 22 kV Northdowns substation. Cable network firmly supplies industrial consumers among the significant potential Harmonic source is Ferroform with 10 MVA load.

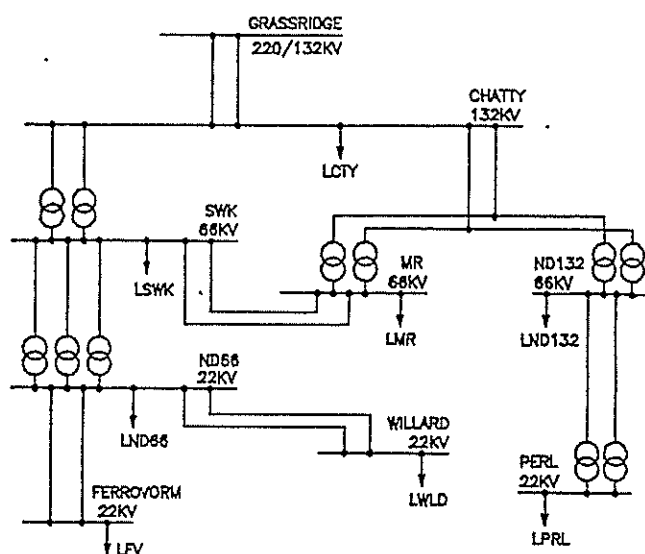


FIGURE 1: PE 132/66/22 kV SIMPLIFIED NETWORK

Ferroform 22 kV substation contains an electric arc-furnaces recognized as a source of harmonics. A combination of arc ignition delays and highly nonlinear arc voltage/current characteristics produce mainly low order harmonics. This is due to continuous motion of melting scrap and the interaction of electromagnetic forces between the arcs on the electrodes during the melting stage. At the refining stage the arc is far better behaved and the distortion is lower. Even though this current does not have a sinusoidal shape, it can still be represented by Fourier series.

Harmonic currents flowing into the system will produce harmonic voltages with each harmonics voltage following the equation:

$$V_n = Z_n * I_n$$

I_n - harmonic current

Z_n - harmonic driving-point impedance of the system

The source of harmonics has been modeled at Ferroform as a current source. The harmonics produced by arc furnaces are superimposed to the existent network as it is shown in Figure 2.

The total Harmonic Distortion is calculated at Ferro-

HARMONIC ANALYSES

CALCULATED HARMONIC DISTORTION AT FERRO

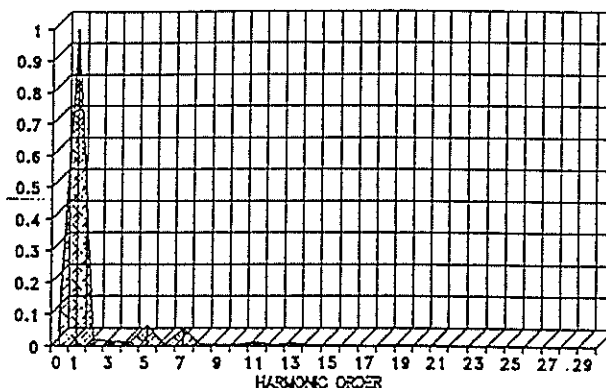


Figure 2: Fourier analyses of voltage at Ferroform

form busbars as the point of common connection (PPC) and at Chatty busbars as the point of supply. The results are presented in Table 1. The voltages and currents in the network are also calculated without the harmonic source as background distortions.

Table 1: THD for one and two sources at Ferroform

CASE	PH SHIFT	VOLTAGE FERRO (V)	VOLTAGE CHATTY (V0)	CURRENT LOAD (I)	CURRENT SOURCE (I)
WITHOUT HARMONIC SOURCE		3.52271	1.79767	3.52217	3.63722
ONE	0	9.8087	4.0877	9.8087	4.102
NONLIN	30	9.2278	4.0007	9.2278	4.10465
SOURCE	60	8.7752	3.2695	8.7752	3.95104
	90	8.60378	2.6577	8.60378	3.85249
TWO	0	18.648	7.65292	18.648	5.39796
NONLIN	30	18.04273	7.0007	18.04273	5.24419
SOURCES	60	17.57933	5.2695	17.57933	4.60478
	90	17.40579	3.6577	17.40579	4.14761
ONE	0	9.8087	4.0877	9.8087	4.102
NONLIN	-30	10.3531	3.3075	10.3531	3.88017
SOURCE	-60	10.73253	1.85742	10.73259	3.59338
	-90	10.86963	1.06141	10.86963	3.51647
TWO	0	18.648	7.65292	18.648	5.39796
NONLIN	-30	19.22491	6.18615	19.22491	4.77543
SOURCES	-60	19.6322	3.15626	19.6322	3.83368
	-90	19.78128	0.86518	19.78128	3.50000

The value of THD at the consumer's busbars of 3.52271% at 22 kV is within the limits in Table 1, while 132 kV value is slightly higher due to the limited modelling.

Harmonic phase angle of the source is varied from -90 to +90 degrees. Theoretically, the angle of each harmonic order can vary from 0 to 360 degrees [3], what is often used in some probability methods of predic-

Table 2: Harmonic Voltage Distortion in % at PPC

	2.3-69kV	69-138kV	>138kV
Maximum for Individual Harmonics	3.0	1.5	1.0
Total Harmonic Distortion	5.0	2.5	1.5

ting of distortions [2]. Some measurements in the real network [1] indicate the range of change to be between 20 and 40 degrees what can be explained by commutant reactance and angle retard of approximately 30° degrees for power rectifiers.

The calculation is performed on KHTP using the Fourier option to analyze voltage and current values at various points of interest. The facility provides waveform values based on a complex amplitude as RMS of sine and cosine coefficient, and "Fractions of fundamentals" for THD for up to 30 harmonic orders. The graphs on Figure 3 and the results in Table 1 show the effect of phase angle variation on THD for load and source currents and voltages.

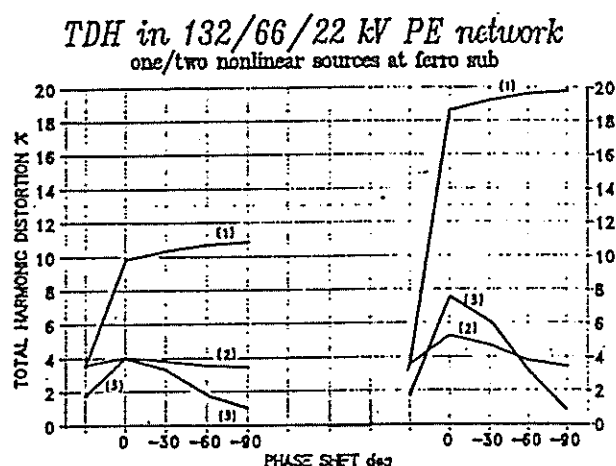
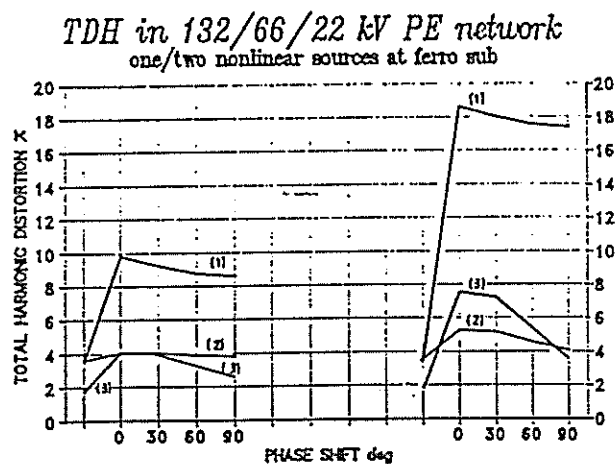


Figure 3: THD related to phase shift from -90 to +90 degree

- (1) voltage at Ferrovarm
- (2) current at Chatty
- (3) voltage at Chatty

The widely accepted figure for harmonic distortion is the quantity of total harmonic distortion THD. It is:

$$THD = \sqrt{\sum I_h^2} / I_f$$

I_h - harmonic component

h - harmonic order

I_f - fundamental component, for power system it is assumed to be power frequency

Harmonic current and voltages are not constant in the network and vary as load conditions changes and as the network parameters are changed. Another source of harmonics of the same amplitude and phase angle is connected to Ferrovarm busbars. The theoretical value of THD will not be twice as much as it is with one source. This also indicates an unpredictable nature of the harmonics. THD is not the only parameter to describe distortions. RMS of the waveform (sinusoidal and cosinusoidal coefficients) are presented in Table 3 and Figure 4.

Table 3: RMS for one and two sources at Ferrovarm

CASE	PH SHIFT	VOLTAGE FERRO (V)	VOLTAGE CHATTY (V0)	CURRENT LOAD (I)	CURRENT SOURCE (I)
WITHOUT HARMONIC SOURCE		0.016377	0.65001	0.016377	0.0017688
ONE NONLIN SOURCE	0	0.01644	0.6504	0.01644	0.001768
	30	0.016438	0.6504	0.016438	0.0017685
	60	0.016432	0.65028	0.016432	0.00176883
	90	0.01643	0.65015	0.01643	0.00176927
TWO NONLIN SOURCES	0	0.016649	0.651879	0.016649	0.00176833
	30	0.016634	0.651686	0.016634	0.00176874
	60	0.016622	0.65089	0.016622	0.00176909
	90	0.016618	0.650339	0.016618	0.00176976
ONE NONLIN SOURCE	0	0.01644	0.6504	0.01644	0.001768
	-30	0.011645	0.650288	0.011645	0.00176807
	-60	0.016459	0.650028	0.016459	0.00176813
	-90	0.016461	0.64993	0.016461	0.00176813
TWO NONLIN SOURCES	0	0.016649	0.651879	0.016649	0.00176833
	-30	0.016664	0.651201	0.016664	0.0017678
	-60	0.016674	0.650249	0.016674	0.00176759
	-90	0.016677	0.649911	0.016677	0.00176826

Harmonic current injected into the system is assumed to have THD of 6.6% THD at Ferrovarm busbars including the characteristics of the existing network is around 9 % and far above the limits.

THD obtains higher values for negative phase shift and this effect must be taken into account when defining THD in the network. RMS values do not change dramatically while the phase angle is varied.

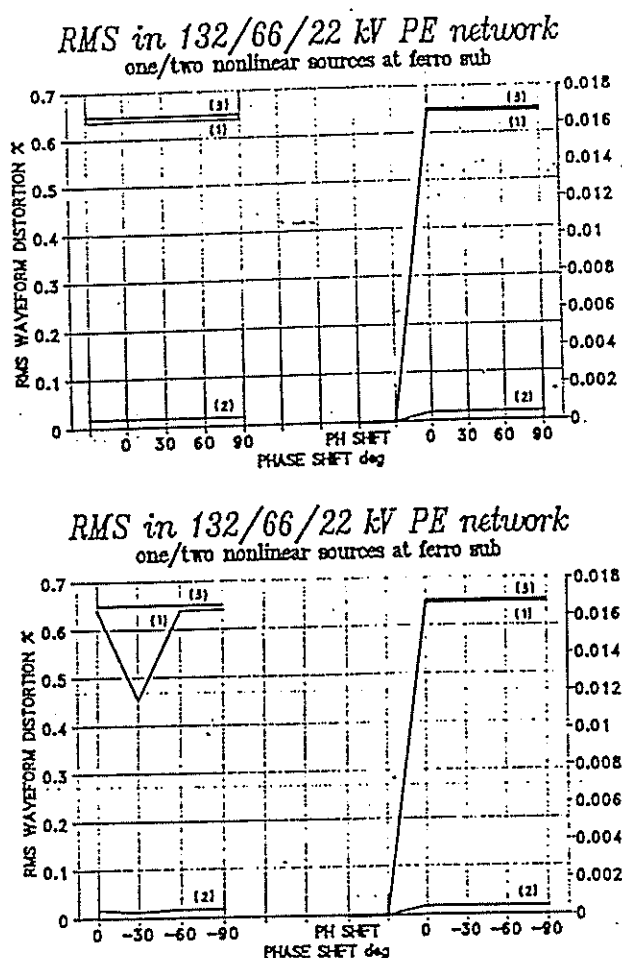


Figure 4. RMS waveform related to phase shift from -90 to +90 degrees
 (1) voltage at Ferrovoorn
 (2) current at Chatty
 (3) voltage at Chatty

3. CONCLUSION

Harmonic load flow in Port Elizabeth network is analyzed determining THD and RMS waveform values at the point of common coupling and at the source. Measurements should justify calculated results and the harmonic source connected to Ferrovoorn busbar should be compensated from the harmonic point of view. This compensation would also take into account the power factor correction.

THD value changes with angle variation, so prediction of the harmonics must embrace this calculation in obtaining highest value. The study should be extended to the rest of 22 kV network in order to obtain the harmonic flow all other consumer PCC's. Capability of the system to absorb the distorted currents depends on a load to short circuit current ratio. In that respect the information on daily load variation should be incorporated into the study.

Optimum compensation improves both, power factor and

harmonic distortion. The devices are preferably connected to the main distribution busbars rather than installed separately, compensating each of the individual consumer. Harmonic study must be a continuous process in the network due to various factors that make nature of harmonics unpredictable.

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BENEFITS OF OPTIMAL POWER FLOW

W. Levy

ESKOM

ABSTRACT - This paper gives an introduction into the capabilities of an optimal power flow program. It also gives some results of an analysis that was carried out using the ESKOM network. The actual state of the network (load, losses, voltages, controls) was recorded for a day, and the analysis compares this state to the state that the network would have been in had an optimal power flow been available. The results show that a saving in losses of approximately 350 MWh could have been achieved over that day with an associated reduction in the number of transformers that required tapping. In addition to the savings from the reduced losses, the decrease in wear and tear on equipment is extremely beneficial.

1. INTRODUCTION

In today's tight economic situation, utilities around the world are trying to exploit their power transmission systems to the fullest, and ESKOM is no exception. This increased utilization of the system can lead to problems such as voltage instability and voltage collapse. Thus utilities are showing interest in methods of controlling voltage levels and dispatching reactive power to achieve economic operation whilst observing system operating limits.

This paper details the analysis of using an optimal power flow (OPF) program on the ESKOM network. The benefits gained by using the program are narrower voltage profiles, reduced losses and a reduced number of control actions.

Section 2 gives a description of the program used, showing its capabilities and the way in which it works. Section 3 then goes on to detail the results of the investigation, comparing the actual system to the system that would exist if the OPF were to be used.

2. CAPABILITIES OF OPF

OPF programs have been designed for both planning and operational use. They are used to determine the best (optimal) manner of operating the electrical power system subject to a set of operating constraints (usually physical limits). Either the active power subsystem or the reactive power subsystem can be controlled, depending on the requirements.

Table 1 : List of SC-OPF controls

Active power controls	Reactive power controls
Generator MW output	Generator voltage/VAr
Phase-shift transformer tap	In-phase transformer tap
DC link MW transfer	Shunt capacitor/reactor
MW transaction	Synchronous condenser
Sheddable load	Static VAr compensator
Switching	Switching

The program which ESKOM is using is called SC-OPF (Security Constrained Optimal Power Flow) [1], [2]. The program works in the following manner - given certain controls in the system which may be used for

optimizing, and applying a set of constraints to the system, an objective function is determined for SC-OPF to optimize the system.

Some of the controls which can be used for optimizing the system are given in Table 1 and the constraints that can be applied to the system are given in Table 2.

Table 2 : List of system constraints

Active power subset	Reactive power subset
MW control range	Voltage/VAr control range
Branch flow	Branch flow
MW interchange	MVAr interchange
Time allowed for control action	Time allowed for control action
Contingencies	Contingencies

The objectives given determine what the optimization does and how it achieves its goal. Two basic objectives are available - the first, for MW scheduling (active power optimization), is the economic dispatch objective which minimizes the cost of MW generation given the cost-curves for each generator. The second, for voltage/VAr scheduling (reactive power optimization), is the loss minimization objective which aims to minimize the MW transmission losses in the system.

By default, the program uses an additional objective of minimum control shift (MCS), which means the program will use as many controls as necessary, but each control will be changed by the smallest amount possible. An alternative is the minimum control action (MCA), which uses as few controls as possible, but each control can be moved by a large amount. This can be graphically illustrated (see Figure 1), where any control is penalized from moving away from its initial point (the dot in the picture). With the MCS objective, a control only starts to be heavily penalized once it has moved away from the initial point, hence many controls can be moved small amounts. With the MCA objective, a control is heavily penalized the moment it moves from its initial point, hence a few controls can be moved larger amounts.

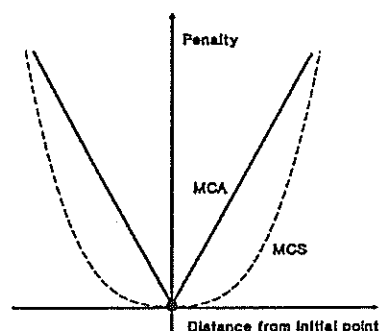


Figure 1 : Difference between minimum control shift (MCS) and minimum control action (MCA)

The ideal situation is to have all controls in the system activated, to have wide ranges for all controls and limits and to have as few constraints as possible - this would then allow the system to be optimized for the greatest benefit. In practice, not all controls are available and certain rigid physical limits are necessary. It must be recognized that from the ideal situation, each control that is made inactive and each constraint that is added, restricts the program a bit more. Thus the right mix of controls, constraints and limits needs to be found, according to practical requirements and the system which is modelled.

3. BENEFITS OF USING SC-OPF

In ESKOM's National Control Centre, there is already a comprehensive suite of programs for active power scheduling, whereas little work has been done in optimizing the reactive power flows. Thus SC-OPF is initially being aimed at the reactive power optimization and the analysis has been carried out to show the benefits of optimally controlling reactive power to minimize losses.

Snapshots were taken from the state estimator every fifteen minutes for twenty-four hours, from midnight to midnight, but only the hourly snapshots were used. The day chosen was a Tuesday in May, thus the loads are typical of a weekday near the winter peak. Results are given comparing the manner in which the system was controlled on that day (the actual, or state estimator data, given by lines with a dot symbol) to the manner in which it could be controlled using an optimal power flow package (given by lines with a plus sign in the results).

3.1. Choosing the objective

The two objectives examined here for optimizing the system are those of loss minimization and minimum control action. Loss minimization is a full-blown optimization in which most of the controls are used to reduce the losses. Minimum control action does not specifically aim to reduce losses, but instead moves as few controls as possible to alleviate any limit violations which may exist.

Ideally, doing a loss minimization at each time step (snapshot) would ensure maximum reduction of losses at all times. This is impractical for two reasons:

- the number of controls that would have to be moved at each step is unacceptable.
- the time taken for the program to do the optimization (typically 5-10 minutes for a loss minimization) plus the time needed to move all the controls, would be infeasible.

Using only the minimum control action objective does not lead to any meaningful reduction in losses. Thus a mix of both objectives is required. For this analysis, a loss minimization was carried out at midnight and thereafter only the minimum control action objective was used. The choice of midnight was chosen because, at that time, the load is fairly static and the operators are not fully occupied as they would be during the day.

Once the controls are scheduled as recommended by the loss minimization, the system should be in a non-violated, optimal state with reduced losses. By the next snapshot (01H00), the load and generation pattern have changed and hence the bus voltages will be different.

Now a minimum control action is performed to fine tune the system and eliminate any possible violations. By the next snapshot (02H00) the load, generation and voltage have once again changed, and a minimum control action is performed. This procedure is carried out for the whole day. Another way of understanding this is that the controls (transformer taps, shunts and generator MVARs) are the variables of the optimization whereas the generation pattern, the load pattern and the bus voltages are determined by the electrical system.

3.2. Losses

Figure 2 shows the real power loss reduction i.e. the optimized losses subtracted from the actual losses. The initial midnight point (00H00) where the loss minimization was done, enables a 16 MW, or 6.3 %, reduction in losses.

Thereafter, using only the minimum control action objective, with few rescheduled controls, the gains are always positive and usually above 10 MW. At light load (up to 06H00), while the load is fairly uniform, the gains are relatively constant. Once the load begins picking up, the gains are somewhat random, determined by the load pattern at that time and by possible network changes. The total reduction of losses for the day, assuming the losses are constant over each hour, amounts to 350 MWh.

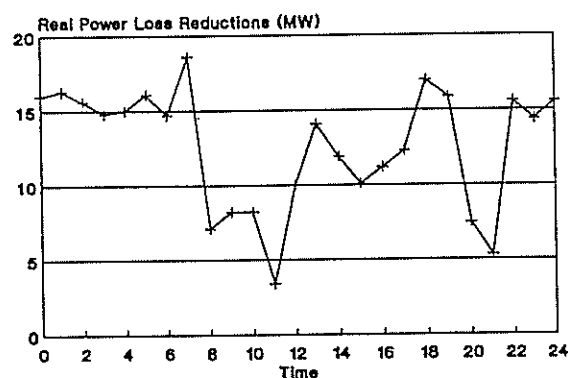


Figure 2 : Real power loss reduction curve

What is important to note is that the initial loss minimization reduces the losses and puts the system into an optimal state. Thereafter, whilst only using the minimum control action objective, the system remains in an "almost optimal" state, with the associated reduction in losses. It was stated earlier that the minimum control action objective eliminates any limit violations - these occur mainly because the voltage limits used by the optimization are narrower than the actual allowed limits. Thus, when a state estimator result is read into the program, certain busses will be flagged as having violated limits.

3.3. Controls

In this section, the control actions are examined that are required to achieve the reduction in losses shown in the previous section. To meaningfully compare the actual control actions to those recommended by the optimization, it is assumed that the state of the system after the loss minimization is the starting point for comparison. Thus the controls for the loss minimization

have not been included at the 00H00 point. Instead another loss minimization was performed at the following midnight (24H00), and these control actions have been included.

3.3.1. Shunts

Figure 3 shows the cumulative number of shunts switched over the day. Until the loss minimization step of 24H00, the number of recommended shunt switchings is almost equal to the actual shunts switched.

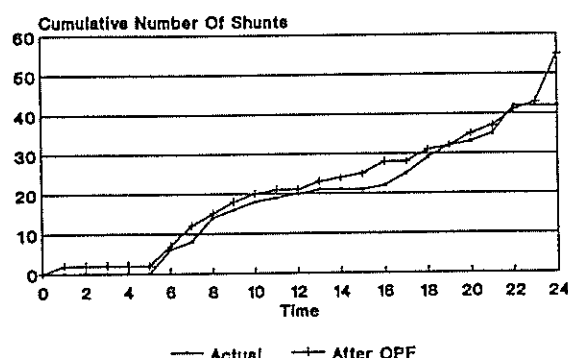


Figure 3 : Comparison of the cumulative number of shunts switched over the day

3.3.2. Transformers

Looking now at transformers, Figure 4 shows the cumulative number of transformers tapped over the day. Until the load picks up at 06H00, the number of transformers that are actually tapped compares well with those recommended to be tapped. From here onwards, the dramatic reduction in tapping can be seen that the optimization offers over the actual system. Even including the loss minimization step of 24H00, the optimization recommends tapping 151 transformers as opposed to the actual system tapping 469 transformers.

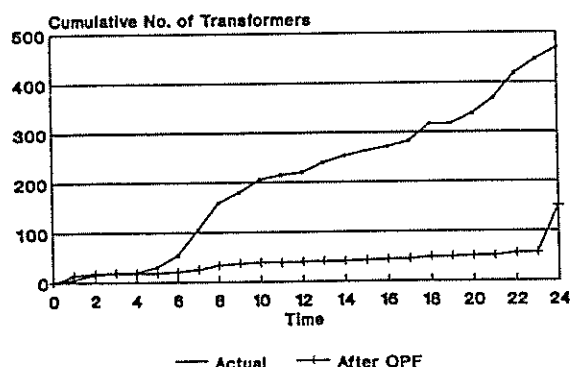


Figure 4 : Comparison of the cumulative number of transformers tapped

3.4. Voltage Profiles

Voltage profiles are given for the 275 kV and the 400 kV subsystems. For the 275 kV voltage profile, all busses with voltages between 279 kV and 280 kV are added

together and shown as the number of busses at 280 kV. All busses with voltages between 280 kV and 281 kV are added together and shown at 281 kV etc. The 400 kV profiles also appear in 1 kV increments.

Figure 5 shows the voltage profile for the 06H00 step which is a sharply increasing load condition. It can be seen that the actual voltages for the 275 kV and 400 kV subsystems have a normal type distribution whereas the voltages after optimization have many of their bus voltages pushed nearer the upper limit.

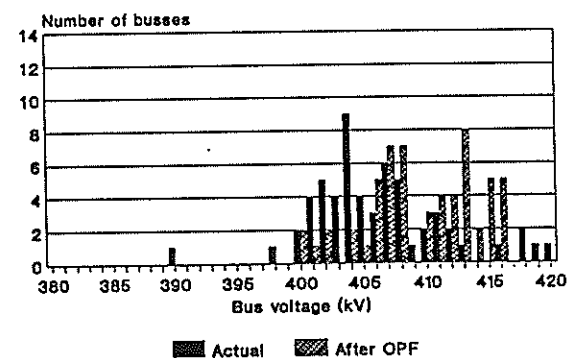
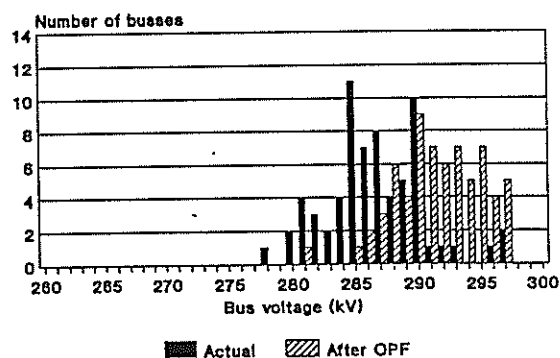


Figure 5 : Comparison of 06H00 voltage profiles

3.5. Contingencies

The scheduled outages are inherently included in the analysis because the state estimator snapshots are used as the data input. If additional outages need to be considered, these are handled via the contingency constraint discussed in section 2. Contingencies are not automatically selected by the program, but need to be decided upon by the user. From a given set of contingencies, the program will automatically rank them in order of severity.

Similar to the method described in section 3.1, an analysis was carried out for the whole day. The start point was also to do the loss minimization at 00H00. Thereafter, for each snapshot, as well as doing a minimum control action optimization, the additional constraint of contingencies was used. This means that in addition to alleviating any violations that may exist, the system must be in such a state that if any of the given outages should occur, the system would still not be violated. Three contingencies were examined, chosen to

specifically be in different geographical areas of the network.

The results show that only a few additional control actions are required to achieve similar loss reductions and lack of violations as the previous analysis. Thus only certain of the results will be shown and briefly discussed.

Figure 6 shows the MW loss reductions achieved by using the contingency constraint. The shape of this curve is different to that without contingencies, but it still shows a positive gain at all times with an average gain of more than 10 MW. The cumulative number of shunt switchings (Figure 7) shows that, until 23H00, the number of required switchings is the same. The 24H00 switchings are those from the loss minimization. Figure 8 shows the dramatic reduction in the required number of transformer tapplings, even including the loss minimization step. The voltage profile of Figure 9 shows that no voltages are violated even though they are pushed nearer to their upper limits.

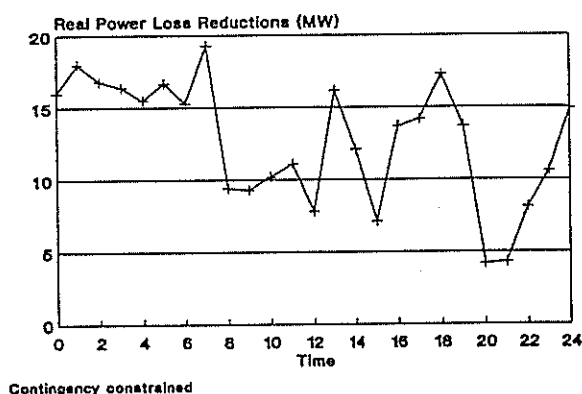


Figure 6 : Real power loss reduction curve for contingency constrained analysis

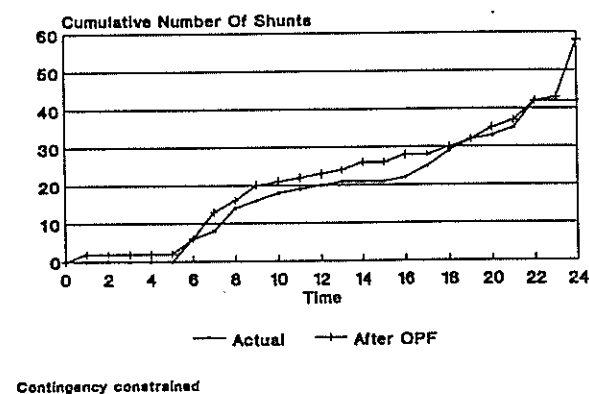


Figure 7 : Comparison of the cumulative number of shunts switched for contingency constrained analysis

4. CONCLUSIONS

This paper has shown a method of optimizing an electrical network which offers considerable benefits. For the ESKOM system, by initially doing a loss minimization, the MW losses can be reduced by more than 6 %. Thereafter, the minimum control action objective suggests only a few control actions to

alleviate any violations, whilst remaining in a "near-optimal" state with on-going reduced losses. Other benefits of doing this are narrower voltage profiles and a large reduction in the amount of transformers required for tapping. Contingencies, in addition to the scheduled outages, may be included so that no violations will occur should that contingency happen.

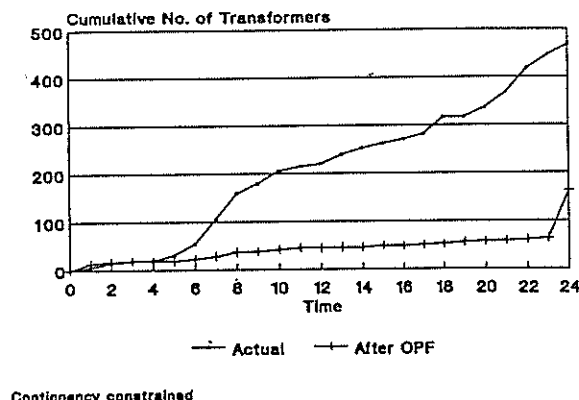


Figure 8 : Comparison of the cumulative number of transformers tapped for contingency constrained analysis

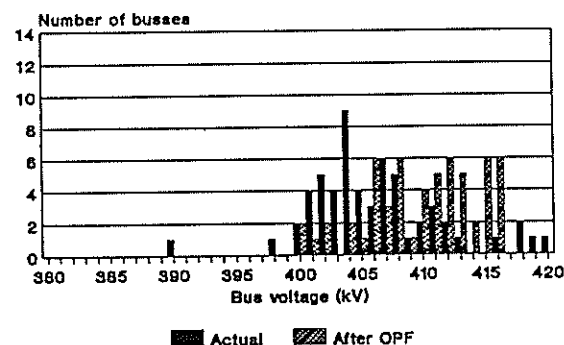


Figure 9 : Comparison of 06H00 voltage profile

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A quantitative examination of negative sequence currents induced in networks due to unbalanced loads.

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Abstract

In this paper, negative sequence currents caused by load unbalances will be examined. The focus of the analysis will be to assess how the unbalance factors vary when load magnitude and unbalance level changes in a radial network. In a meshed network the effect of load position and multiple unbalanced loads will be investigated.

1 Introduction

In a perfectly balanced network, all phase currents are equal and only positive sequence currents are present. If any network element is unbalanced, negative and zero sequence currents will be induced in the network. The cause of this unbalance could be untransposed lines, and unequal phase loads. In this paper, only the effect of load unbalances will be examined, as they are the most common type of unbalance found in a power system.

Unbalanced loads, will lead to a phase asymmetry in currents in the whole network. This asymmetry can be expressed as a ratio of the negative sequence component to the positive sequence component, or an *unbalance factor*. It is used for voltages and currents.

From a supply point of view, unbalance factors, at nodes, have to be determined, to ensure that no other consumer, will suffer a loss in the quality of supply due to the unbalanced load. One of the problems associated with negative sequence current, is that rotating machinery develop heating problems. This is due to rotating machines having a lower negative sequence than positive sequence impedance. Hence for even a small voltage unbalance at the terminals of a machine, a large negative sequence current will flow[1] and eventually cause the insulation of the machine to be degraded. To prevent this, manufacturers, have quantified the voltage unbalance factor, that can be tolerated at the terminals of the machine, to be at most 2%[2]. The utility thus has to ensure that the voltage unbalance factor is less than two percent for

buses where there are rotating machines. In South Africa the voltage unbalance factor has been set to 1.5%[3].

In this paper, the load unbalances will be examined using a balanced radial and meshed network. The transformers and loads will be connected in star-delta and delta respectively, to prevent the introduction of zero sequence current. The network diagrams are figures 12 and 13 in the appendix. The unbalanced three phase loads were modeled in the following way. On phase B the magnitude of the load would equal P . On phase A and phase C, the load would equal $P - \delta P$ and $P + \delta P$. In this study, δ ranged between 5 and 33. Hence when reference is made to a 10% unbalance level, it implies that the ratio of the load across the three phases are 23.33 : 33.33 : 43.33.

2 Radial network example

This network has three nodes with a generator, transmission line, transformer and one unbalanced load. The first investigation determined the influence of the load magnitude and load unbalance level on the network unbalance factors in a network. In order to do this, the load magnitude was increased and the unbalance level was kept constant. Next, the load unbalance level was varied and the load magnitude was held constant. The emphasis will be placed on how voltage asymmetry varies at the terminals of the generator. The results for the cases of load magnitude and load unbalance level are depicted in figures 1 to 3 and 4 to 6 respectively.

2.1 A varying load magnitude with a constant unbalance level.

As the load magnitude increases, the profile of the negative sequence voltage increases. This is due to the phase voltage differences between the three phases of the system increasing as the load size increases. At node 1, the unbalance factor for voltages ranges from 1.413% to 4.036% and this generator will experience problems. The highest negative

sequence voltage, ranging from 2.995% to 9.439% occurs where the unbalanced load is connected. Hence the voltage distortion with an increasing load magnitude cannot be ignored.

The negative sequence current unbalance factor ranges from 17.76% to 19.29% as the load magnitude increases. These values are very close, and it is the result of the current always dividing in a 23.33 : 33.33 : 43.33 ratio for a 10% load unbalance. The amount of negative sequence current thus depends on the size of the load unbalance and not the load magnitude.

The active and reactive power losses with an unbalanced load increases quite significantly as the load magnitude changes. The reactive power loss increases steeply in relation to the active power loss. However, this is due to the load magnitude and not to the size of the unbalance. A balanced load with these magnitudes exhibits a similar graphical relationship between active and reactive power loss, as in our case. The only difference caused by a load unbalance is that the losses are slightly higher than in the case for the same magnitude of balanced load. Hence losses are not a critical factor here.

2.2 A varying load unbalance level with a constant load magnitude.

The negative sequence voltages increase linearly at each node for an increase in the unbalance level at the load bus. The magnitude of the negative sequence voltage decreases, as the electrical distance from the point of unbalance in a radial network increases. The 2% unbalance factor for voltages is exceeded when the load unbalance level is at 10%. The negative sequence voltage at the generator terminals range from 2.55% to 10.80%.

The unbalance factors for the negative sequence

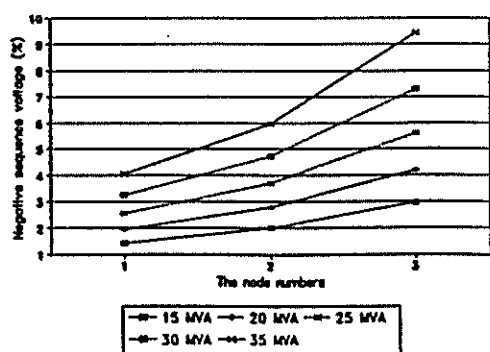


Figure 1: A profile of negative sequence voltage for varying load magnitudes and an unbalance level of 10%.

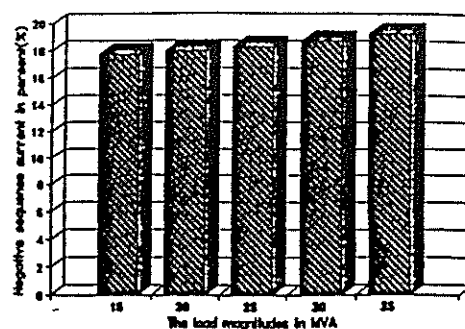


Figure 2: A profile of negative sequence current for varying load magnitudes and an unbalance level of 10%.

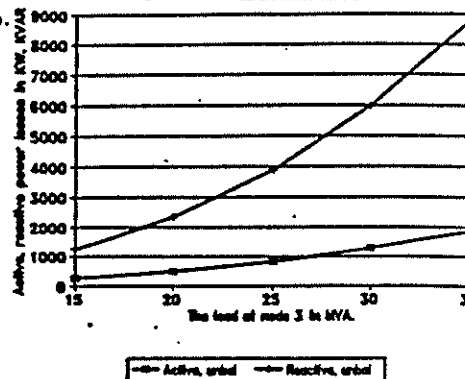


Figure 3: Active and reactive losses for varying load magnitudes and an unbalance level of 10%.

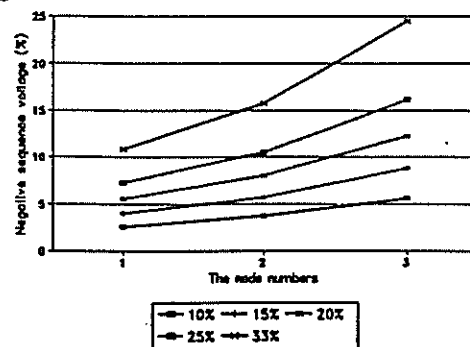


Figure 4: A profile of negative sequence voltages for varying unbalance levels and a constant load magnitude.

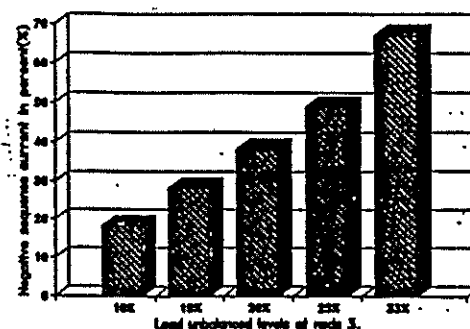


Figure 5: A profile of negative sequence current for varying unbalance levels and a constant load magnitude.

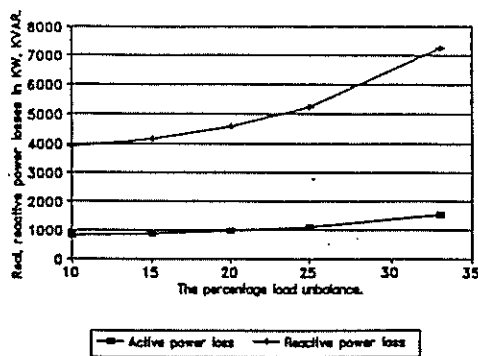


Figure 6: Active and reactive losses for varying unbalance levels and a constant magnitude.

current increases linearly as the level of load unbalance increases. This is due to the current distribution between the phases increasing proportionally to the unbalance level. The values range between 18.30% to 67.21%.

The active and reactive power losses increases significantly as the level of load unbalance increases. The reactive power loss increased by 86% for a load unbalance increase from 10% to 33%.

The results of the two cases shows that the magnitude of an unbalanced load introduces a definite amount of voltage and current distortion and that this distortion becomes worse as the load unbalance level increases.

3 Meshed network example

This network has six nodes with two generators, five transmission lines, two transformers and three loads. The first investigation was aimed at determining the distortion introduced when a load, of constant magnitude and unbalance level, is placed at the three different nodes sequentially. The other two nodes would be balanced. The results are depicted in figures 7 to 9.

The second investigation examined the effect of different levels of unbalance at a particular node. The results are illustrated in figures 10 to 11.

The last case determined how multiple load unbalances distort voltages and current in the network. These results are illustrated in figures 7 to 8.

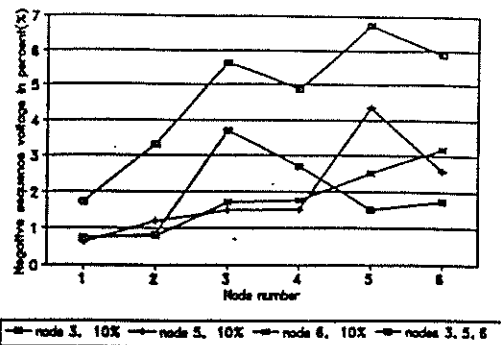


Figure 7: A profile of negative sequence voltage for a constant load unbalance level and magnitude at nodes 3,5 and 6.

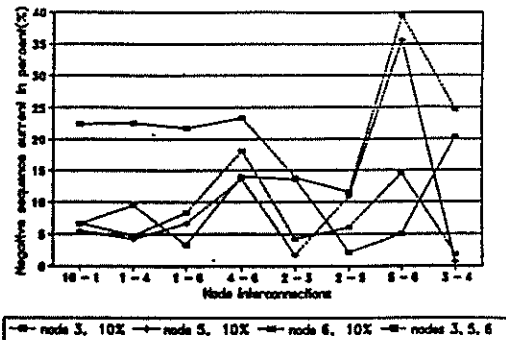


Figure 8: A profile of negative sequence current for a constant load unbalance level and magnitude at nodes 3,5 and 6.

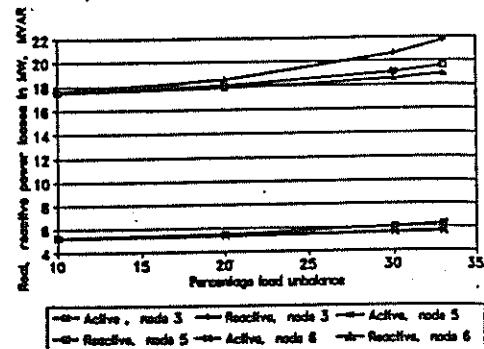


Figure 9: Active and reactive power losses for a constant load unbalance level and magnitude at nodes 3,5 and 6.

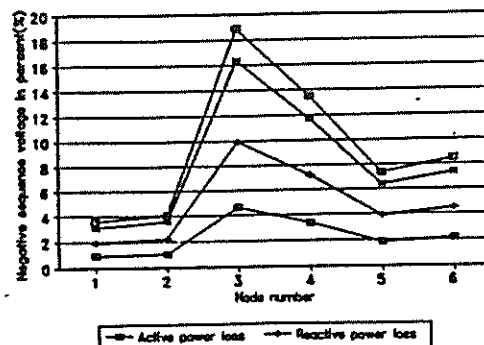


Figure 10: A profile of negative sequence voltage for varying levels of load unbalance at node 3.

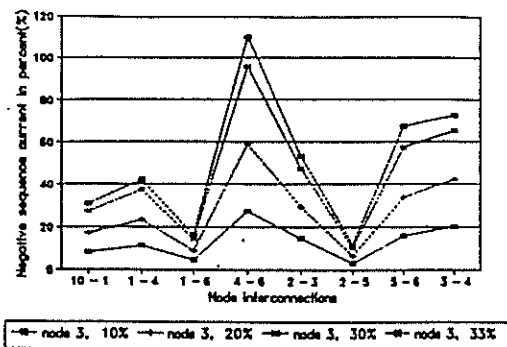


Figure 11: A profile of negative sequence current for varying levels of load unbalance at node 3.

3.1 Constant load magnitudes and unbalance levels at different nodes.

The negative sequence voltage profile peaks at the point where the unbalanced load is connected. The unbalance factors for voltages at nodes 1 and 2, the generator terminals, are less than 0.80% and 1.2% respectively. The voltage profiles differ quite markedly from each other and this indicates, that the electrical distance affects the level of voltage distortion introduced in the network.

Similarly, the widely varying negative sequence current profile, shows how electrical position can affect the flow of current. The peak sequence current occurs in the branches incident to the unbalanced load.

The active and reactive power losses for this case differs very slightly. The active losses range from 5.27 to 5.29 MW for the three cases. The reactive power losses range from 17.94 to 18.01 MVar. These figures are similar since the total system load does not change and the power flow within each branch is equal. This seems to indicate that the power losses are sensitive to the total system load and not the electrical position of that load.

3.2 A varying load unbalance level with a constant load magnitude at node 3.

As the unbalance level increases, the negative sequence voltage induced in the network increases. A 20% load unbalance level, at node 3, results in a 2% unbalance factor at nodes 1 and 2. In comparison to the radial example, this difference in the size of the unbalance level, is due to the different MVA ratings of the machines.

The voltage distortion decreases as the electrical

distance between nodes increases. This is reflected in the fact that the voltage distortion at node 5 is less than at node 6.

The negative sequence current profile increases almost linearly due to an increase in load unbalance. The largest distortions in current are in the branches incident to the unbalance node. The unbalance factor peaks on the line connecting nodes 4 and 6. This is a reflection of the fact that power flows from node 4 to node 6. The active and reactive power loss also shows an increase as the load unbalance level increases.

3.3 Three nodes with equally unbalanced loads and magnitudes.

The interaction of three loads, simultaneously unbalanced by 10% on phase A, is non-linear. From fig. 7, the fourth voltage profile shows that one cannot simply add the negative sequence voltages of individual unbalanced loads, to obtain the resulting profile. The sequence level at generator one is barely tolerable at 1.73% whereas at generator two, the limits for the machine have been exceeded at 3.32%. The fourth negative sequence current profile, in fig. 8, shows a large increase in the amount of current unbalance present. Both voltage and current distortions are consistently larger than the individual negative sequence profiles. The peak voltage and current distortions occurs at nodes 5 and 6 and on the line between these nodes. This shows that a multiply unbalanced network experiences more distortion than when it is unbalanced by one load only. In contrast, though, a single load with a large unbalance level could produce more distortion than the case where these three nodes are unbalanced by 10%.

The active power losses were 5 MW and the reactive power losses were 16.73 MVAR. It would appear that the total active and reactive power losses decrease when a constant unbalanced load is placed simultaneously at three nodes than when a constant unbalanced load is placed at only one node.

4 Conclusion

An attempt was made to determine how the aspects of load magnitude, electrical distance and unbalance level affects the amount of negative sequence current that can be injected into a particular network. The results are grouped according to network type and are as follows:

• Radial Network

1. For the network studied, the different load magnitudes did not affect the unbalance factor for negative sequence current. A load of 15MVA or 25MVA will produce nearly the same amount of negative sequence current in a network if each is unbalanced by 10%.
2. For different levels of unbalance, the unbalance factor for negative sequence current increases as the level of unbalance increases.
3. For the cases 1 and 2 above, the unbalance factor for negative sequence voltage increases as the load magnitude and the level of unbalance increases.
4. For the two cases of load magnitude and load unbalance level, the voltage unbalance factor generally exceeded the 2% level.
5. The active and reactive power losses increases as the load unbalance level and magnitude of the load increases.

• Meshed network

1. The peak negative sequence voltage occurs at the unbalanced load node.
2. The peak negative sequence current occurs in the branches incident to the unbalanced node.
3. The worst case, with three loads unbalanced produced a voltage unbalance of 1.73% and 3.32% at generators one and two respectively.
4. The voltage and current distortion due to load unbalances at three different nodes, is non-linear.
5. The active and reactive power losses increases if the level of unbalance at a particular node increases.
6. The active and reactive power losses for the unbalanced loads at the three different positions were nearly equal.

Possible areas for future work would include the investigation of the interaction of unbalanced loads in order to reduce the negative sequence voltage profile. This would mean examining how phase rearrangements could help to reduce the unbalance factors in the network. More strenuous unbalanced load conditions such as a mixture of single and double phase loads should also be examined.

5 References

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6 Appendix

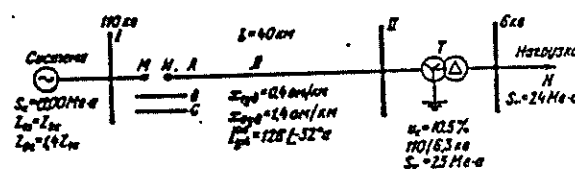


Figure 12: Schematic diagram of the radial network.[4]

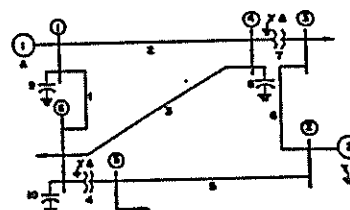


Fig. R.2. A 6-node network (79).

TABLE OF IMPEDANCES IN PU ON A 100 MVA BASE (TABLE 11)

TYPE	CONNECTING NODES	IMPEDANCE Z1 = Z2	IMPEDANCE Z0
GEN 1	1	10.240	11000
GEN 2	2	10.480	10.032
LINE 1	1-6	11.036	12.064
LINE 2	1-4	10.740	11.850
LINE 3	4-6	10.814	12.060
LINE 4	2-6	11.280	13.840
LINE 5	2-3	12.100	15.280
TR 1	6-4	10.600	10.600
TR 2	3-4	10.288	10.288

Figure 13: Schematic diagram of the meshed network.[5]

APPLICATIONS OF PARAMETER PLANE TECHNIQUE IN POWER SYSTEM DYNAMIC STUDIES

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The parameter plane methods provide a powerful method for studying system stability alongwith information about the desired parameter combination for a system to satisfy a required performance. However, it is noted that undergraduate students are not conversant with these techniques as these are rarely mentioned in modern textbooks on control theory and not included in the relevant syllabus. The aim of this paper is to highlight the importance of these methods to complement classical well-known techniques. This paper focuses on the practical applications of the above-mentioned methods to stability, optimisation and sensitivity studies in power systems.

1. INTRODUCTION

Parameter methods (parameter-plane and parameter-matrix formulations) were developed to solve high-order problems too laborious to solve using classical techniques [1-6]. The philosophy of the methods is to map contours in the p-plane onto the plane of two design parameters, using the characteristic equation as the mapping function. The domain decomposition technique is an alternative to Siljak's generalisation of the Mitrovic's method [2]. These techniques are extremely powerful for the study of asymptotic stability in linearised autonomous systems. They are particularly valuable for determining the controlling parameters which meet a given system performance of root requirement. However, it is noted that they have not been widely applied to power system studies. The aim of this tutorial-type paper is to highlight the importance and practical applications of these methods to complement classical techniques based on stability criteria of Nyquist, Routh and Hurwitz and the design methods of Bode, Nichols and Evans of classical control theory.

Consequently, this paper focuses on the practical applications of the above-mentioned methods to stability, optimisation and sensitivity studies in power systems. Parameter plane techniques are used here for the study of subsynchronous resonance. Results are presented to demonstrate the effect of several controller parameters on the dynamic stability of a series capacitor compensated transmission power system. Detailed sensitivity analysis is presented for a system with a specified degree of absolute stability and with a specified degree of relative stability.

The techniques are also applied to the characteristic equation of a general optimal control system showing their usefulness in the study of the effect of system parameters on the optimal transient performance. The use of these methods as computer-aided procedures for multivariable control system analysis is presented.

2. PARAMETER PLANE TECHNIQUE

This technique also known as the Domain Separation (D-Decomposition) is a very powerful tool for stability studies. The method allows for the determination of the conditions of system stability; it also shows how to select appropriate parameter values in order to obtain better system response as it can be used to study the degree of damping. The importance of the technique is clearly reflected in Russian literature, where it has been presented in textbooks [4,5]. However, the method has apparently attracted little interest among authors of control theory textbooks at the undergraduate level.

The parameter plane technique is based upon a graphical procedure for factorising characteristic polynomials in the plane of two or more parameters. Characteristic curves are plotted by mapping certain contours from the p plane onto the parameter plane of interest. These curves provide information on relative damping, natural frequency and pole locations. The method has been applied to linear-continuous, non-linear, sampled-data, and optimal linear systems [7]. The basis of the technique is that the characteristic equation

$$\Phi(p) = \sum_{i=0}^n a_i p^i = 0 \quad (1)$$

of the system under investigation is regrouped in such a way that the whole of the parameter space can be decomposed into a number of domains, for each of which there is a fixed root distribution with a known number of roots with positive real parts and a known number with negative real parts [6]. Knowing the root distribution in any one domain, the distributions in all other domains can be written down by inspection, and those domains corresponding to stable conditions identified.

The regrouping can be done as follows:

$$\Phi(p) = \sum_{i=0}^n a_i p^i = \sum_h b_h p^h + P_1 \sum_k c_k p^k + P_2 \sum_m d_m p^m = 0 \quad (2)$$

where P_1 and P_2 are the two real parameters of interest which appear linearly and separably in the characteristic equation.

Equation (2) can be rewritten as

$$Q(p)P_1 + N(p)P_2 + R(p) = 0 \quad (3)$$

where $Q(p)$, $N(p)$ and $R(p)$ are polynomials in p . By substituting $p = jw$ and separating real and imaginary parts equation (3) can be written as

$$Q_1(p)P_1 + N_1(p)P_2 + R_1(p) = 0 \quad (4)$$

and

$$Q_2(p)P_1 + N_2(p)P_2 + R_2(p) = 0 \quad (5)$$

$$P_1 = [R_2(p)N_1(p) - R_1(p)N_2(p)] / \Delta \quad (6)$$

$$P_2 = [Q_2(p)R_1(p) - Q_1(p)R_2(p)] / \Delta \quad (7)$$

$$\Delta = [Q_1(p)N_2(p) - Q_2(p)N_1(p)] \quad (8)$$

The transformation from the p plane to the P_1P_2 plane through the mapping function (equation 1) must be nonsingular, i.e. $\Delta \neq 0$, $\forall w \geq 0$.

By assigning suitable values of w from 0 to ∞ , it is possible to find sets of values for P_1 and P_2 for which the system is stable for zero damping. Such points may be plotted as a gain locus in the P_1 and P_2 plane. The gain locus divides the gain plane into two domains: stable and unstable. When w increases, the locus is shaded on the left-hand side if $\Delta > 0$ and viceversa. The shaded region correspond to that of the stable operating zone [2]. The Routh-Hurwitz Criteria are then applied to points on each side of the stability limit boundary to confirm the stable and unstable zones.

3. APPLICATIONS

3.1 Analysis of Subsynchronous Resonance

Series capacitor compensation provides a direct approach to increase the transmission capabilities

which is often the most economical solution. However, it has been established that self-excited oscillations are likely to occur, under certain conditions, in a series capacitor compensated transmission system. Excitation control has been studied as one of the measures to suppress these oscillations. In this section, the parameter plane technique is utilised to demonstrate its effectiveness to study the problem of subsynchronous resonance. The power system under investigation consists of a dual-excited synchronous machine connected to an infinite bus through a compensated transmission line (Figure 1).

To study the system dynamic performance, a small displacement model of the system is developed. Excitation control, governor action and the shaft torsional dynamics are included in the model. Such a model can be written in matrix form as

$$[A] \cdot [\Delta x] = 0 \quad (9)$$

The elements of matrix $[A]$ and the system parameters are defined in reference [8].

The objective is to investigate how different excitation parameters (T_e , K_{oI} , K_{1I} , and K_{2I}) affect the system stability. In investigations for the unregulated system (no excitation control) reveals that the operating point corresponding to armature resistance $R = 0.6$ p.u., $X_C = 0.6$ p.u. and load angle $\delta = 20^\circ$ is unstable. It is of interest to plot the stability zone in the T_e/K_{oI} plane to demonstrate the effects of these excitation parameters on the stability zone (Figure 2).

Further analysis shows that the incorporation of additional signals in the form of the first and second derivatives (K_{1I} and K_{2I}) of the excitation regulator helps to suppress low frequency self excitation oscillations (Figure 3 : curve a).

Since the size of the stability zone gives a measure of the reserve of stability, the gains of the excitation regulator can be properly selected such that the flexibility of setting the regulator gains is maximised [9].

Transient stability limit can be determined by employing the well-known step-by-step method or other methods like the Runge-Kutta. In order to study transient response, the Nyquist plot is often used. This is not very convenient to use for studying systems with controllers working on derivative feedback signals as the Nyquist plot corresponds to only one set of values of the system operating conditions. Thus, a number of such plots are required for different values of the derivative gains.

The locus determined in Figure 2 (or Figure 3 : curve a) defines the absolute value of stability since the imaginary axis of the root plane is mapped onto the parameter plane. Any other line may be mapped so that the enclosed domains correspond to an arbitrarily chosen degree of damping by substituting $p = -\lambda + jw$ into equation (3) and repeating the procedure as before. Figure 3 (curves b,c) shows the curves

corresponding to different degrees of damping, λ . Controller parameter values can thus be selected in order to ensure the maximum degree of damping as the amount of damping present is a measure of the quality of transient response.

3.2 Optimum Controller for Automatic Generation Control (A.G.C.)

In this section, the practical application of the parameter plane technique to stability, optimisation and sensitivity studies for an A.G.C. system [10] is reported.

The characteristic equation for a two area system provided with integral controllers can be obtained as

$$a_0 p^6 + a_1 p^5 + \dots + a_5 p + a_6 = 0 \quad (10)$$

It is found that equation (10) contains non-linear combination of the two parameters of interest (K_I and B). Equation (10) can however be rearranged so that $1/K_I$ and B appear linearly, thus permitting the application of the parameter plane technique. Stability zones (for different values of λ) are obtained in the plane of $1/K_I$ and B , and then transformed to the K_I and B plane (Figure 5). The best values of K_I and B are then chosen as those corresponding to stable zone $\lambda_{\max}$.

An optimum controller setting based on a suitable cost index is obtained through minimum quadratic integral error criterion.

The dynamic responses (Figure 6) computed for the two cases (corresponding to $\lambda_{\max}$ and the optimum control) are found to be similar

4. CONCLUSIONS

The importance of the parameter plane technique as a tool to supplement existing classical control schemes has been highlighted through two examples in power systems.

It is proposed that this technique be introduced to senior-level students.

5. LIST OF SYMBOLS

a, b	coefficients
h, i, k, l, m, n	integers
P_1, P_2	physical parameters
$\Phi, M, N, Q, L, R, \Delta$	polynomial functions
p	complex variable ($p = -\lambda + j\omega$)
$T_e, K_{ol}, K_{II}, \text{ and } K_{2I}$	current regulator settings
K_I	integral controller gain
B	frequency bias parameter

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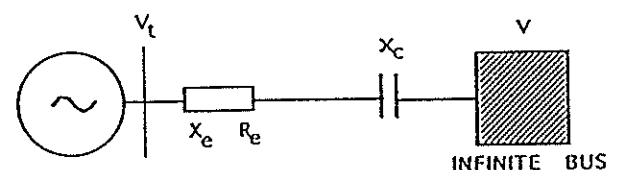


Figure 1: System Investigated in Section 3.1

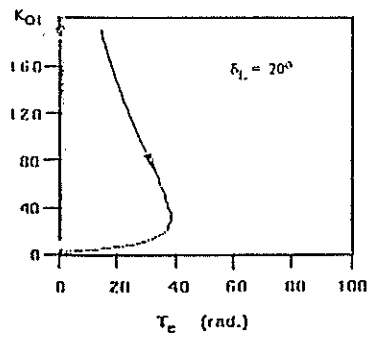


Figure 2: Stability zones in the T_e - K_{01} plane

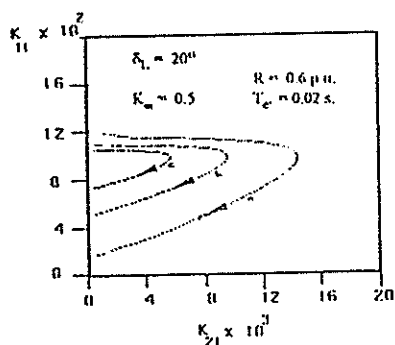


Figure 3: Stability zones in the K_{21} - K_{11} plane

	a	b	c
λ	0.0	0.002	0.004

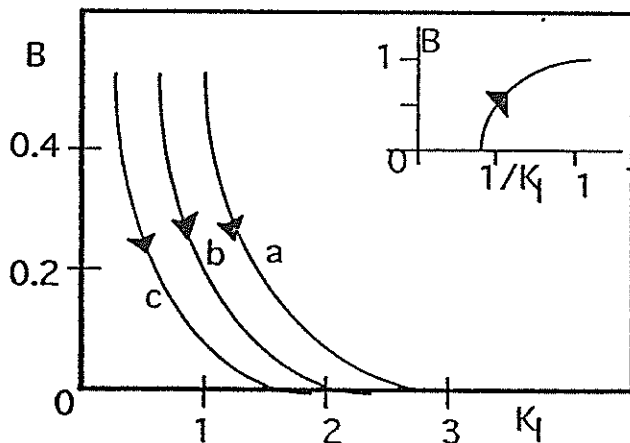


Figure 4: Stability zones in

(i) $1/K_I$ - B plane

(ii) K_I - B plane

	a	b	c
λ	0.0	0.10	0.2

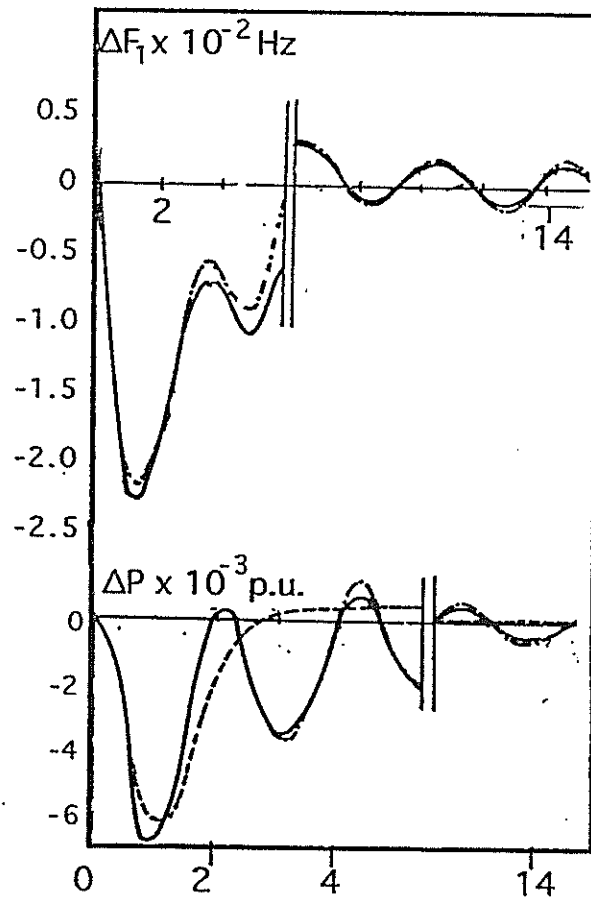


Figure 5: System dynamic responses

— $\lambda_{\max}$
 - - - ISE

THE PROTECTION OF A POWER TRANSMISSION NETWORK USING PHASOR ANGLE INFORMATION

C Cairns

Eskom (Technology Research and Investigation)

Phasor Angles across transmission lines can now be economically measured with the advent of satellite disseminated precise-time. These angles give the state of the power transmission system relatively simply and reliably with significant implications for power monitoring, control and protection practices.

The power flows in a power transmission system are determined by the dynamical relationships between all elements of a power system. These elements include the power generation, the load demand, the state of the system circuit breakers, the reactive compensation applied on various lines and many other factors which together influence the state of the system to some greater or lesser extent. Currently the overall stability of the system can fall prey to a protection philosophy which is based on a compromise which encompasses every possible fault in order to protect the system against it. There are some combinations of system condition which, while they do not threaten system stability, nevertheless trigger local protection to the detriment of the system as a whole.

For some time now it has been proposed that a form of advanced protection be developed which would react optimally in a local sense to global events occurring in the power transmission network. This protection has been termed 'adaptive protection' and has been defined as follows:

"Adaptive Protection is a protection philosophy which permits and seeks to make adjustments to various protection functions in order to make them more attuned to prevailing power system conditions"[4]

The major problems associated with this form of protection is that it requires an accurate, timeous assessment of the current system state and that while protection based on immediate (local) conditions may not be optimum, it is guaranteed to protect the system from damage whereas protection based on a global assessment of the situation might conceivably sacrifice some of the network as a consequence of some unforeseen sequence of events. These two issues can be decoupled from each other and will be addressed separately.

State assessment

The state of the system is generally taken as the vector V of bus voltage magnitudes and the vector θ of voltage angles which can be deduced from the phasor angles measured simultaneously throughout the system, arbitrarily assigning one of the busbars as an absolute reference. Previously it was not possible to measure these angles directly and they were calculated from the measured power flows in the system using techniques which suffered from the inherent nonlinearity of the power-voltage relationship, measurement inaccuracy and parameter variation. The state estimator operated by Eskom National Control for example requires between 5 and 10 minutes to estimate the network state - often erroneously. Direct measurement of the phasor angles eliminates the interim power-to-voltage conversion and speeds up the process of determining the system state thereby improving the system observation and control bandwidth.

Phasor angles can be measured directly by a number of techniques, all of which require the measurement events to be synchronised in the spatially distributed power transmission system. This synchronism requirement can be seen in one of the simpler procedures for measuring voltage angles:

Assume two remote counters located at either end of a transmission line. These independently count to 360 every 20 ms and are reset and started every time a zero crossing of the 50 Hz voltage signal is detected. Assume now that clocks operating synchronously at either busbar stop these counters simultaneously. The difference between the two counters then indicates the phasor angle between the two busses. Referring to Figure 1 the phasor angle is given by the difference between Counter #1 and Counter #2. Several techniques for synchronising the local clocks on the busbars have been developed, all of which fall into one of two categories:

1. Clocks are periodically moved from their deployed positions to a master clock which is used to synchronise them. This technique requires very stable clocks which are capable of maintaining independent time within specification between synchronisation. The clocks tend to be expensive and the system becomes prohibitively expensive once the expense of the infrastructure and transport costs

required to maintain it have been factored in.

2. Communication links between the clocks are used to remotely synchronise them. The problem with this method is that it is necessary to determine the path propagation delay across the link in order to compensate for it when issuing synchronisation signals. Radio links are susceptible to continually varying propagation delays as the conditions along the signal path change and multipath effects can also give rise to varying delay. Telephone links do not have sufficient bandwidth to communicate synchronisation signals effectively and are also often routed via radio links. Other band limited techniques such as differential phase calibration are also not practical due to telephone switching practices.

More recently it has become possible to access the time keeping system used by the Navstar Global Positioning System (GPS) and to use the time disseminated by the satellites for synchronising remote clocks. Navigation using GPS is dependant on maintaining a common time standard between satellites and navigation receivers and a number of commercially available navigation receivers issue a 1 pulse-per-second signal which is synchronised within 300 ns of true GPS time. This signal can be used to synchronise a local clock reliably and relatively inexpensively with significant gains in system simplicity:

1. Inexpensive free-running local clocks can be used as they are unlikely to drift significantly during the 1 second interval between resynchronisation. This also makes the system relatively immune to oscillator ageing effects.
2. Each GPS based clock operates independently of all others. It can be remotely deployed and operated without any other support infrastructure or calibration requirements.
3. The local clock maintains synchronism within 1 μ s. This performance is significantly better than that obtainable using radio or telephone synchronisation.

Eskom Technology Research and Investigation (TRI) have developed a measurement instrument which uses this method to measure the voltage angle. This instrument is based on a PC platform which is used

to perform local data processing and to relay information to a central processor via a modem link. Referring to Figure 2 and considering the signal paths through the instrument, a positive sequence voltage signal derived from the distance protection on a network busbar is buffered and fed to a 50 Hz bandpass filter which eliminates noise and harmonics. The conditioned signal is then the input to a positive zero crossing detector which is simply the combination of a comparator which detects when the voltage signal performs a zero crossing, and a monostable which performs the dual functions of a divide-by-two (to eliminate one of the zero crossings) and the issuing of a clear pulse to the angle counter. The angle counter is reset and cleared on every positive zero crossing event by the positive zero crossing detector. The counter then counts until it is latched by a signal derived from the GPS 1 pulse-per-second signal. The synchronously measured angle in degrees is then given by the following relationship:

$$\text{Angle} = \frac{\text{Angle count}}{\text{Max count}} \times 360^\circ$$

Where *Max count* is the maximum count achievable during one 50 Hz cycle (20 ms) and its input is the output of an oscillator which is phase locked to the 50 Hz signal. The oscillator signal is divided down to generate a suitable clock for the angle counter.

This instrument has demonstrated a measurement accuracy of better than 2 degrees which is considered acceptable for state estimation[1]. It has been postulated that network protection based on phasor angle information alone might be feasible and this possibility will be examined by Eskom.

Protection based on state assessment

Power system stability can be improved by changing the network to accommodate and counter undesired changes in the network. Applied properly actions such as generator dropping, load shedding and switched reactance compensation change the network with a consequent improvement in network stability. Various strategies and tactics are used as a basis for actions such as 'islanding' whereby the network is split into a number of independent networks which are separately stable or alternatively selected loads may be severed from the network as required.

Fundamental to all of these stratagems are measures which firstly determine whether the network is approaching instability and secondly the location of the instability if present. The voltage angles across the busbars turn out to be very good indicators of the system stability as they are highly reactive to

changes in the network - much more so than the busbar voltage magnitudes which are the other state measure (or the currents which are also measured). A very early indication of incipient instability can be gained by measuring the rate of change of the phase angles in the network - the faster they change the more serious the condition. Furthermore the phasor angle across the line is a direct indication of its stability - the larger the angle, the closer the line is to instability[2].

Protection based on phase angles therefore promises to be superior to the standard protection which is based on current and voltage measurements. It has been proposed that the Cape region might be protected using some form of adaptive protection based on phasor angle information and this possibility is currently being studied. If adaptive protection of the region proves feasible it will be necessary to develop a dedicated protection strategy and tactics to accommodate the structure of the Cape's unique system. It has been proposed that a neural network might be used to process phasor angle information to yield information that can be used to protect the system. If so, it is possible that a generic solution to protecting power transmission networks may be at hand.

Conclusion

Previously it has not been possible to measure the state of a large power transmission system directly because it was not possible to synchronise

measurements economically and reliably. Protection and operational practices were therefore based on current and voltage measures and the state of the system was determined from these measures with some difficulty. In the future, as phasor angle measures are performed on more of the power system, the measured state of the system will approach its true state and this information will then become the basis for most of the operational practices on the system. Eskom is currently investigating the use of this information to improve the protection of the power transmission system in the Cape.

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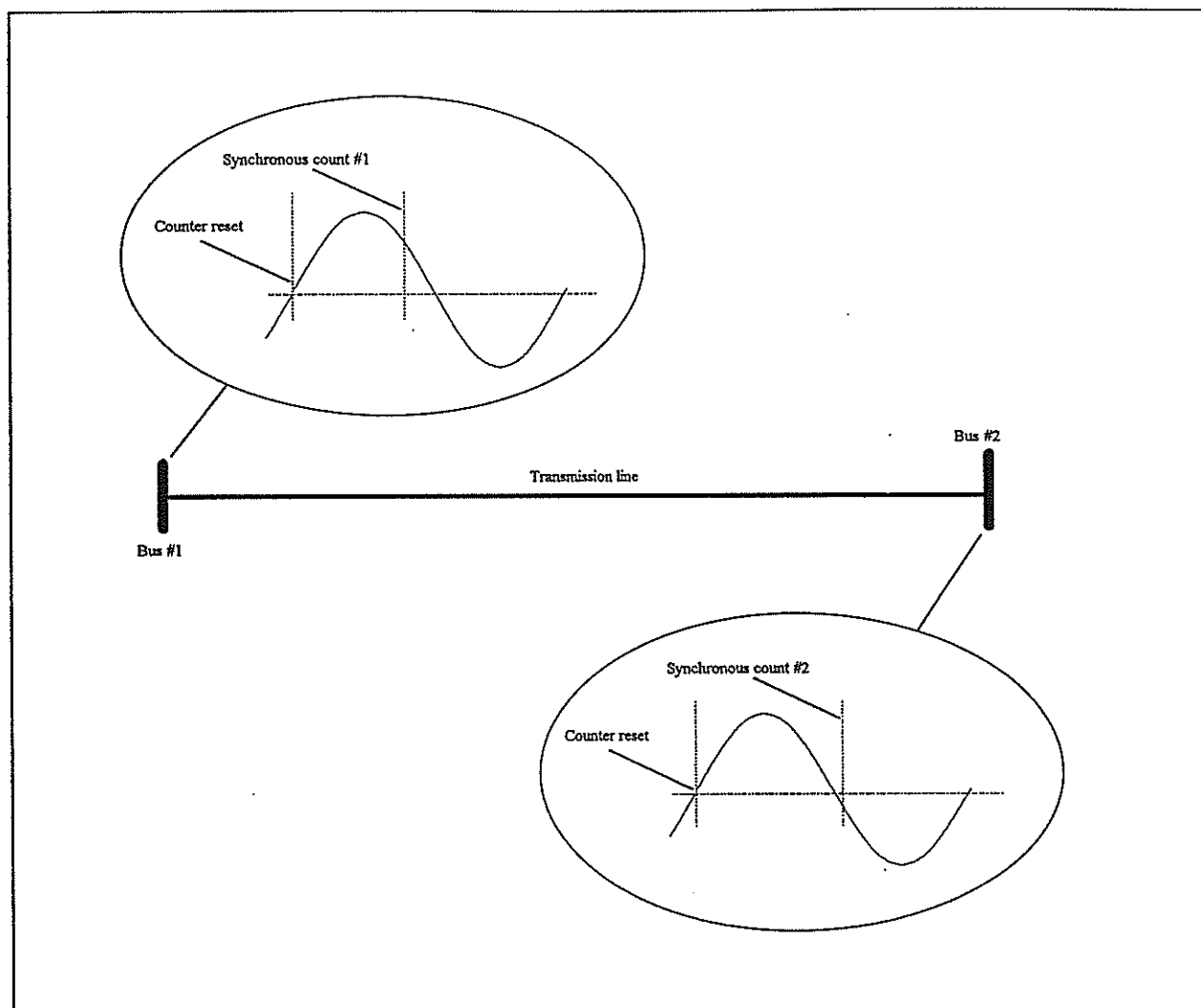


Figure 1: Synchronous measurement of phasor angle between busbars

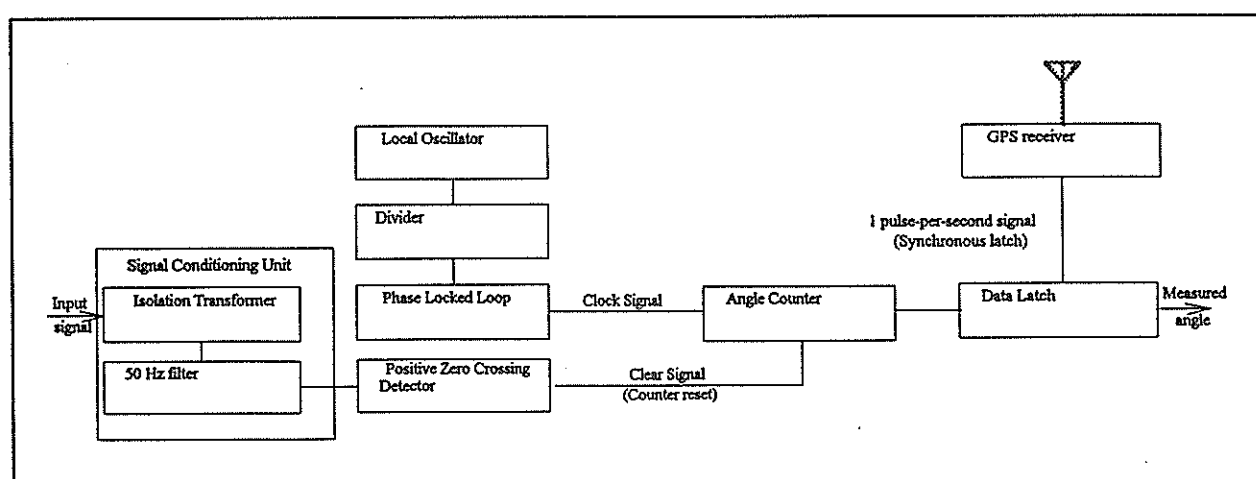


Figure 2: Functional modules of the Voltage Angle Measurement instrument

THE IMPORTANCE OF CONSIDERING PROTECTION PERFORMANCE IN POWER SYSTEM STABILITY STUDIES.

J van Eyssen

ESKOM

This paper investigates protection behaviour and performance during conditions of power system instability. Mathematical models for distance protection are introduced in the power system model when doing stability studies. Transient and dynamic stability studies are presented to investigate the behaviour of distance protection. By including protection models in transient and dynamic stability studies, it is shown that a cascade effect of switching lines may occur.

1. INTRODUCTION

A loss of synchronism between systems will affect the relaying on these systems. Protection relays such as distance relays may detect an out-of-step swing and may therefore operate to trip their breakers. This gives reason for concerned about protection in a routine stability study. Protection operations beyond those responding directly to a fault may aggravate a stability problem.

Two cases are presented. Firstly, a transient stability study is performed, and secondly a dynamic stability study is performed. Both these cases serve to show the influence that unstable conditions may have on distance protection behaviour.

2. CONSIDERING PROTECTION MODELS IN STABILITY STUDIES

The network used for the purpose of the study, can be seen in Figure 1. Generator data is listed in Table 1. The system consists of nine buses, three generators and three loads. This network is a benchmark used for comparative studies in the power system community.

The nine-bus network is a simple system. Complicated power plant such as SVC, series and shunt compensation, motor loads etc., are not present in this network. Due to its simplicity, only generators, transmission lines and loads need to be modelled. For the study, each generator was represented by a two axis generator model. The model represents a round rotor synchronous machine. The model has a field circuit, one damper winding in the direct axis and two damper windings in the quadrature axis. Transmission lines

were modelled by their π equivalent circuits, and loads were modelled as constant impedance.

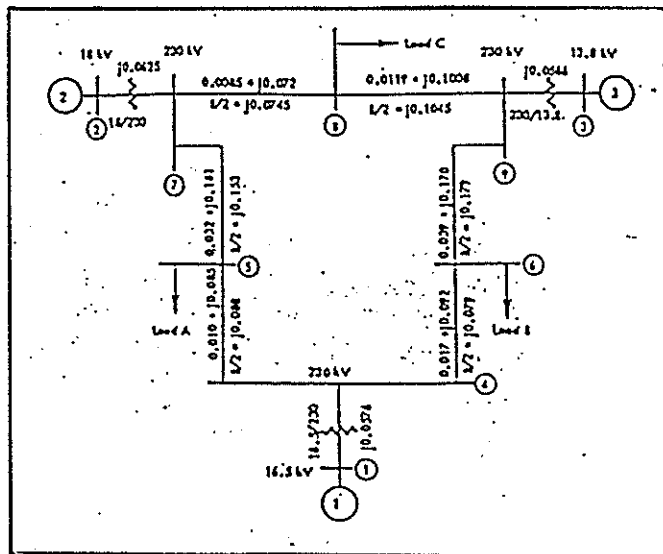


Figure 1: The 9-bus network

Table 1: 9-bus generator data

Generator	1	2	3
Inertia H	6.396	3.302	2.3828
Speed damping D	0.00	0.00	0.00
X_d	1.581	1.651	1.680
X_q	1.531	1.59	1.610
X_d'	0.380	0.232	0.232
X_q'	0.955	0.380	0.320
X_d''	0.252	0.171	0.171
X_q''	0.248	0.171	0.171
X_1	0.291	0.102	0.095
T_{d0}	5.390	5.9	5.890
T_{q0}	1.500	0.535	0.600
T_{d0}'	0.053	0.033	0.034
T_{q0}'	0.135	0.078	0.080
$S(1.0)$	0.0905	0.105	0.121
$S(1.2)$	0.345	0.477	0.610

In this study, transient and dynamic stability studies are performed to investigate the behaviour of protection. Transient stability studies are aimed to determine if the system will remain stable when it is subjected to severe disturbances. Severe disturbances are transmission line faults, sudden load changes, loss of generation etc. For the study, a three phase transmission line fault is simulated close to a generator. Due to the severity of the fault, synchronism is lost within 1 sec. (See Figure 2)

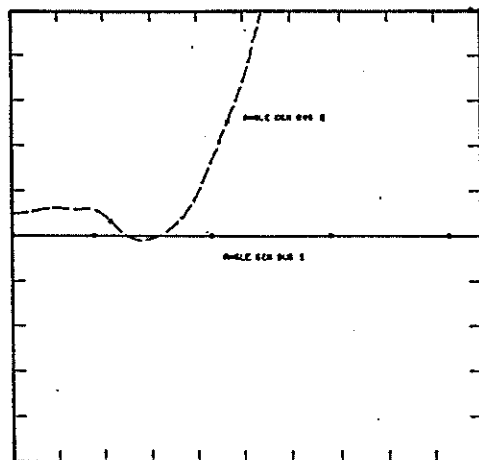


Figure 2: Transient instability.

Dynamic stability studies investigate system behaviour when the system undergoes slow or gradual changes such as a small load change, line switching etc. A power system is dynamically stable if it returns to its original state after a small disturbance. To illustrate dynamic instability, a small load change is introduced on the system. (See Figure 3.)

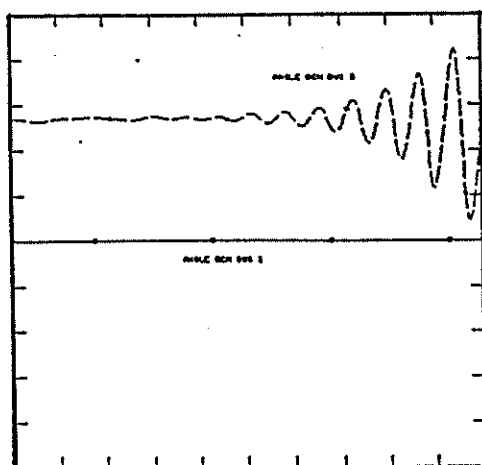


Figure 3: Dynamic instability.

In addition to the 9-bus network model, distance protection was modelled. The distance relay model is a general model for distance relays having three time delayed mho characteristics, with an offset mho characteristic for zone 3. The impedance characteristic can be seen in Figure 4. Distance relays were modelled to protect every line in the 9-bus network.

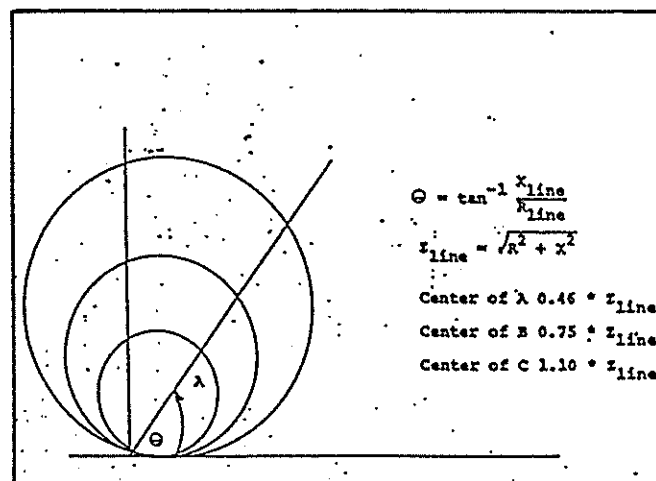


Figure 4: Distance relay model.

To investigate distance protection behaviour when performing transient and dynamic stability studies, two cases are presented:

CASE 1: Transient stability

A permanent three phase fault is simulated on line 6-9 close to generator 3. Distance relays are located on all the lines. The distance relays on line 6-9 clear the fault. Distance protection performance for the system can be seen in Table 2.

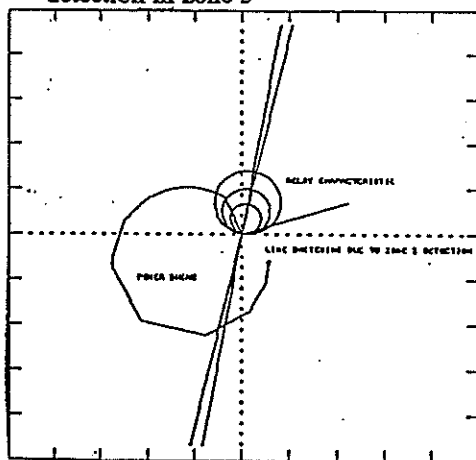
Due to the fault, relays close to the fault location detect impedance inside the operating zones. This behaviour is expected because these relays are backup relays for the fault on line 6-9.

After the fault is cleared, system oscillations starts and the system becomes transiently unstable. For the period after the fault is cleared, distance relays detect impedance entering their operating zones. Distance relays on lines 5-7 and 4-5 detect impedance inside their zone 1 characteristics. (See Figure 5.) These relays switch lines 5-7 and 4-5 respectively.

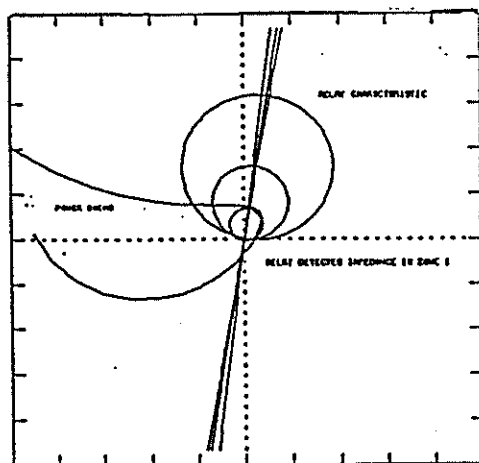
Table 2 : Distance relay performance for case 1.

Relay Location	Relay Performance During Fault	Relay Performance After Fault Clearance
5-4	n	z1 z2 z3 O
4-5	n	z1 z2 z3 O
4-6	z3	n
6-4	n	n
9-6	z1 z2 z3 O	n
6-9	z2 z3	n
9-6	n	z1 z2 z3
8-9	z2 z3	z1 z2 z3
7-8	z2	z2 z3
8-7	n	z2 z3
7-5	n	z1 z2 z3 O
5-7	n	z1 z2 z3 O

n = no detection O = relay operation
 z1 = detection in zone 1 z2 = detection in zone 2
 z3 = detection in zone 3



Relay on line 57



Relay on line 45

Figure 5: Relay characteristics for relays located on lines 5-7 and 4-5.

CASE 2: Dynamic stability

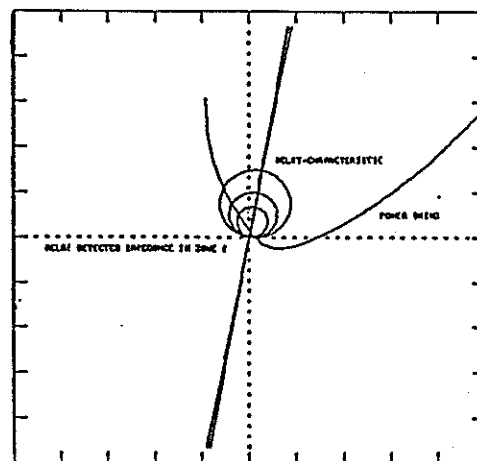
A small change in load is introduced on the system. The system becomes dynamically unstable and is characterised by growing oscillations. (See Figure 3.)

Relays all over the system detect impedance inside their characteristic zones. (See Table 3.) Relays located on lines 4-5 and 4-6 detect impedance inside their zone 1 characteristic and lines 4-5 and 4-6 are switched. The impedance seen by these relays is shown in Figure 6.

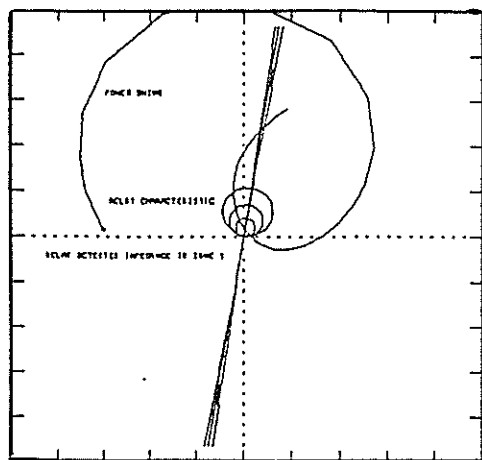
Table 3 : Distance relay performance for case 2.

Relay Location	Relay Performance During Fault
5-4	z1 z2 z3 O
4-5	z1 z2 z3
4-6	z1 z2 z3
6-4	z1 z2 z3 O
9-6	z1 z2 z3 O
6-9	z2 z3
9-6	z2 z3
9-8	n
8-9	z3
7-8	n
8-7	n
7-5	z2 z3
5-7	z2 z3

n = no detection O = relay operation
 z1 = detection in zone 1 z2 = detection in zone 2
 z3 = detection in zone 3



Relay on line 45



Relay on line 4-6

Figure 6: Relay characteristics for relays located on lines 4-5 and 4-6.

3. CONCLUSIONS

In this paper, a transient, and a dynamic stability study were presented. The stability studies were performed on a 9-bus benchmark network. By including distance protection models in the power system model, it was shown that distance relay performance can be influenced during conditions of transient or dynamic instability. The simulated disturbance caused the system to experience power oscillations. Due to these oscillations, impedance entered into relay characteristic zones. A cascade effect of line switching, due to relay action, occurred.

These results show the importance of including relay models when doing stability studies. To prevent the misoperation of distance relay protection during conditions of instability on a power system, the possibility of cascade switching of lines, due to protection performance, must be kept in mind.

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Relay Test Station Developed

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ABSTRACT

The software used in this application is the Electromagnetic Transients Program (EMTP). This program has the ability to model network components through distributed parameters.

This feature allows the user to monitor attenuation and propagation of signals (transients) at any point in a network, resulting from a fault at any point selected in the network.

INTRODUCTION

A Power system transient computer program, namely, the Electromagnetic Transients Program (EMTP), can now be used for the commissioning and maintenance of new and operating relays in distribution systems.

This program will allow the user to enter the characteristics of an existing distribution system so that a full simulation on the operation of the system, at any node, may be observed. The parameters of the system may also be altered in order to create a worst case scenario required for testing equipment.

The EMTP is capable of simulating electrical equipment in fine detail given the required input data. This fine detail may be required when using the program to perform pseudo tests on sensitive electrical equipment.

PROCEDURE

An interface has been developed for the EMTP in order for the program to be used as a tool for simulated in-circuit testing of relays.

Simulation test sets controlled by personal computers were introduced in about 1980 [1]. This greatly improved the consistency in test routines. When using the EMTP to generate the digital signal required for conversion into analog format, the accurate large scale test equipment is no longer required.

The required voltages, inductances and capacitances are modeled by the EMTP with sufficient capacity in order to generate the requirements of the interface unit.

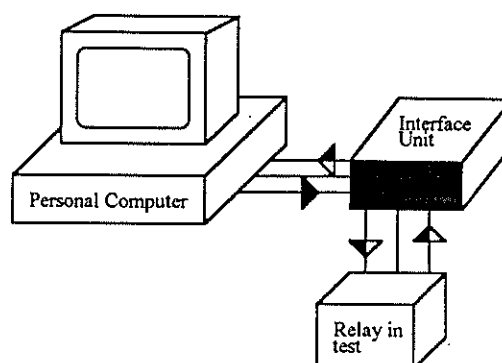


Figure (1) Hardware interface unit

As illustrated in figure (1), the relay would be connected to a personal computer via an interface module. The point at which the relay should be inserted into the power system is chosen and the simulation package is set up accordingly. A few well chosen test cases (faults) are entered into the system via the EMTP and all the required tests are carried out.

The response curves and dependent variables are recorded for future analysis. In this way new relays may be commissioned and tested in circuit through simulation.

A set transient signal can be used to determine the state (condition) of the relay through comparison of initial responses to new responses. This will present an idea of the service periods for relays in the various network stages. The change in responses also analyzed may be due to heat, chemical contamination and abnormal voltages. [2]

A windows based program has been developed to aid this test procedure. As illustrated in figure 2. the package comprises the following modules:

2a. The preprocessor module

This module allows the user to insert all the data required by the EMTP for a complete simulation of any system or transient function.

This module also offers features like : accessing the recorded data then plotting the curves against a set of chosen axes, printing the curves as file copies for reference or reports and creating new libraries for recorded output.

2b. The Graphics analysis module

This module will provide the user with the facility of plotting up to three curves simultaneously and thereby allowing the user to analyze the transient fed to the relay as well as the voltage and current responses captured by the hardware interface unit.

2c. The record library module

A copy of all the requested input and output curves may be stored in this library for convenience and ease of access.

2d. The control module for the Hardware interface unit

A system of algorithms provides the user with sufficient control to govern the sampling rates and other features offered by the system.

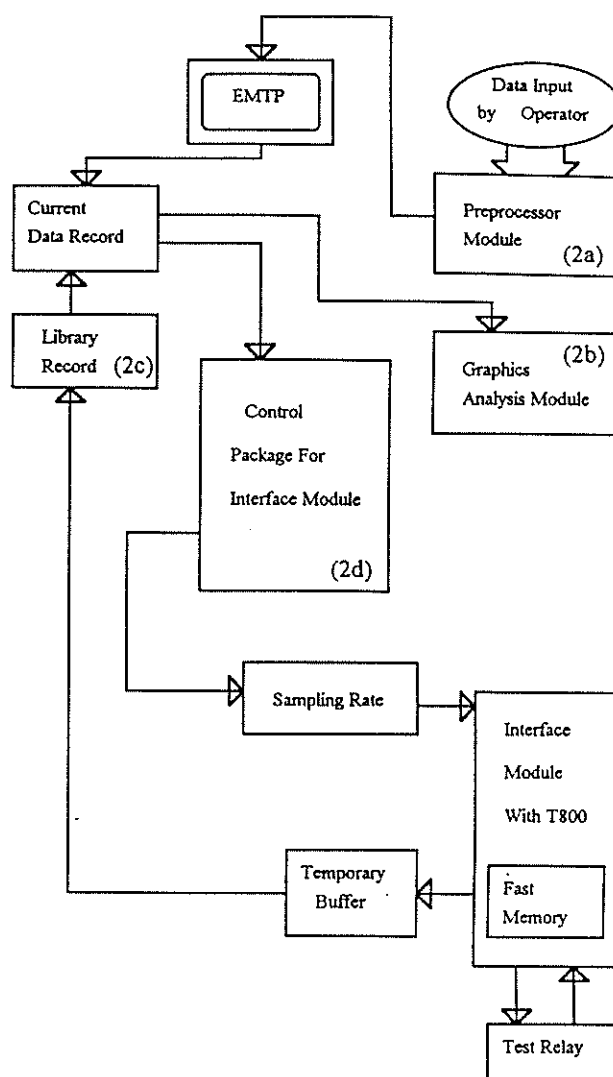


Figure (2) Windows Based Package

The user-friendly package which has been developed in order to facilitate the inexperienced operator, while offering the facility of a relay library where all the test cases may be stored as a means of comparison. The complete package was written for the Windows version 3.1 software package, but is also available in it's DOS format.

The interface module used as indicated in figure(1), makes use of the T800 transputer which features a bi-directional data transmission rate of up to 2.3 MBytes/sec.

The T800 is programmed in ANSI C. Although the T800 transputer may not be the fastest, it is sufficient for performing the required task

A Technique employed by the relay test module built, is to make use of stepped sine wave testing of protective devices. With stepped sine wave testing, the changes from prefault to fault conditions occur at some distinct point in time.[3,4]

All of the voltage and currents will change simultaneously.

A number of microprocessor based relay models have been investigated for incorporation into the EMTP during 1992.[5]

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A COMPUTER-BASED WAVEFORM SYNTHESIZER FOR PROTECTION RELAY TEST APPLICATIONS

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ABSTRACT

Power system faults can give rise to highly complex transient voltage and current waveforms, particularly during the periods immediately following the faults. Modern microprocessor-based protection relays can in principle respond intelligently to such input signals. This gives rise to the need for test equipment with the capability to evaluate protection relay performance for representative waveform data recorded for actual field conditions, or obtained alternatively by simulation using software packages such as the Electromagnetic Transients Program (EMTP). This paper describes a computer-based, three-phase, arbitrary waveform synthesizer for protection relay test applications.

1. INTRODUCTION

The lack of operating specifications for relays have, in the past forced utilities to rely on relay manufacturers for performance information on their relays. New relay testers have recently been developed by these utilities to test all kinds of protection. Among these are the Real Time Digital Simulator (RTDS) [2], Programmable Transmission Line (PTL) [1,4] and the RElay Testing EXpert (RETEX) [3] systems. Many of these systems integrate the 3 basic building blocks of relay testing, namely :

- Waveform data acquisition, for example an EMTP simulation or a fault recording.
- Waveform synthesizing via a digital-to-analog (D/A) converter and voltage and transconductance amplifiers.
- Interpretation of the recorded results.

Figure 1 shows the basic block diagram of a complete test system. This paper deals specifically with the D/A converter, although the amplifiers and the test data handling are also discussed.

Although the system was developed for relay-testing, it is already being used in other fields such as machine drives, signal generation for parameter estimation techniques and to drive an Epstein-frame used for iron-loss measurements.

2. SYSTEM OVERVIEW

The D/A card developed for the system is transputer-based, which means all parameters are set either in the transputer software or by the user at runtime.

The waveforms generated by the D/A converter are amplified by linear, dc-coupled voltage- and transconductance amplifiers. The output signals from

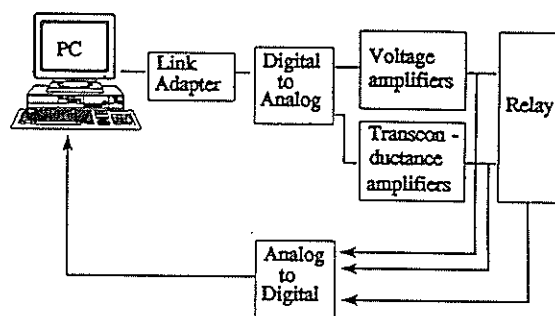


Figure 1 : A one-line block diagram of the complete relay testing system.

the amplifiers are compatible with the output specifications of standard potential and current transformers with respect to amplitude and frequency bandwidth.

An EMTP software interface was developed to convert a multiphase EMTP simulation into a multiphase test signal, making it possible to test all the relays simultaneously for a simulated multiphase fault. Any recorded or mathematically generated sequence of points can, however be used as a data block in a test sequence.

3. HARDWARE

The hardware, as seen in Figure 1, consists of a link adapter (explained in the next section), a D/A converter, transconductance and voltage amplifiers, an analog to digital converter and a personal computer.

3.1. Link adapter.

The link adapter is a plug-in card that facilitates serial communication between the personal computer and the transputer on the digital to analog converter module. Communication takes place at 10 Mbits/sec.

3.2. The digital to analog (D/A) converter

The D/A converter module is a dedicated transputer-based unit, which is configured and controlled via the link adapter from the host computer. The unit features an onboard transputer, 32 kwords of read-only memory (ROM), 32 kwords of random access memory (RAM) and two 12-bit, four-channel D/A converters. Figure 2 shows a block diagram of the D/A converter module.

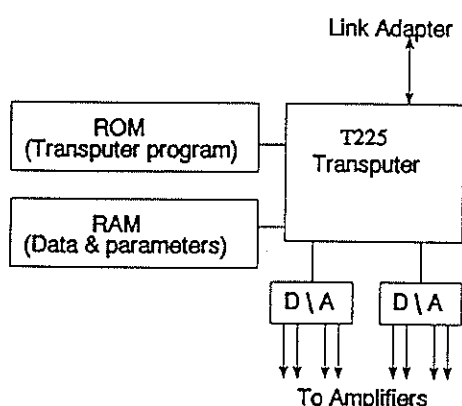


Figure 2 : Block diagram of the D/A converter module.

The test data is downloaded to the D/A converter's RAM before the testing takes place. The data conversion rate is programmable at configuration time and implemented by dividing the 5 MHz frequency of the onboard oscillator. Any frequency between 77 Hz and 100 kHz can be obtained with at least 99% accuracy, whereas the minimum accuracy from 100 kHz to 250 kHz is 97%. Two additional timer chips on the transputer board can allow the frequency to be brought down to several days if they are cascaded, should the D/A application require such a frequency. Note that 8 different output frequencies can be implemented consecutively according to the requirements of each block of data.

The OCCAM code for the transputer routines can be stored on the ROM of the D/A converter card, leaving the 32 kwords RAM free for data storage. Otherwise the OCCAM code could be downloaded to the onboard RAM along with the test data at runtime. This code typically occupies 7 kwords of RAM, leaving 25 kwords available for test. This means that a minimum of 25 kwords of test data points can be accommodated

by the on-board RAM.

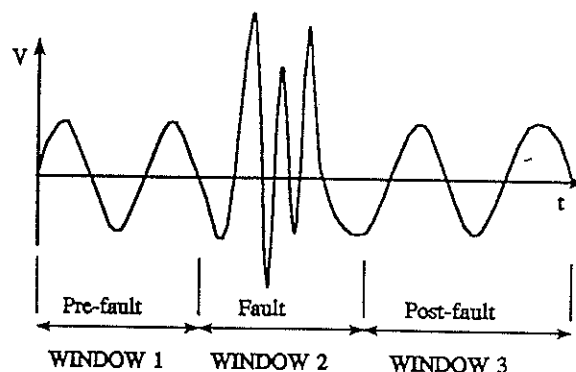


Figure 3 : Typical use of one test channel.

A typical relay test sequence might consist of a pre-fault, a fault and a post-fault stage as shown in Figure 3. The pre-fault and post-fault conditions would typically be repetitive waves. Only one cycle of each of these waves has to be described in the test data. This cycle will then be repeated by the D/A conversion process for as many times as specified by the user. The fault or transient condition, however, would typically be described as a one-shot, high-frequency waveform at a much higher sampling rate than that of the pre-fault and post-fault conditions.

Each of the three time-slots in Figure 3 will be referred to in this paper as a *window*. The 3 windows showed in Figure 3 can be extended to include a total of 8 windows, each of which has its own output frequency. Figure 3 shows the output of 1 out of a possible 8 channels. Typically, 7 channels would be used to test protection systems, namely 3 voltage and 4 current channels.

The speed at which the D/A conversion can take place makes the use of output filters unnecessary for protection applications.

The digital-to-analog converter is programmable by means of data passed to it from a Pascal program. The Pascal program prompts the user for the following information :

- NUMBER OF ACTIVE OUTPUT CHANNELS (maximum is 8) :
- NUMBER OF DATA BLOCKS (WINDOWS) PER CHANNEL (maximum is 8):

Each window is associated with a block of waveform

data, and the user is prompted for the following information for each :

Window 1 (e.g.)

- Number of samples : In the case of a cyclic wave, this will be the number of points that will constitute one cycle of the wave. In the case of a non-cyclic wave as obtained by an EMTP transient simulation this will be the number of simulation data points that results in satisfactory resolution.
- Sample frequency : This specifies the conversion rate for the data points in this window. For example, a cyclic window with a base cycle consisting of 100 points and a required wave frequency of 50 Hz will require a sampling frequency of 5000.
- Number of cycles (0 for an infinite number of cycles): A 0 (zero) in this field will result in the repeated D/A conversion of the data in the window until triggered manually (at runtime) by the user. Otherwise the conversion will proceed to the next window after completion of the specified number of cycles.

There exists a trade-off between the D/A converter's maximum conversion rate and its versatility. For an application that requires the maximum sampling rate of 250 kHz, only one output channel can be utilized and parameters like the number of cycles has to be programmed into the OCCAM code, making the system less versatile at runtime, but flexible in the sense that all changes are still made in the software. This means that the same D/A converter card can drive 8 output channels, each converting 8 unique sets of data in succession, at 32 kHz by running one PC-program -- and then converting a set of data at 250 kHz only by running another PC-program.

Two programs were developed - one that provides maximum versatility (8 channels, 8 data-windows each) - and one that provides minimum flexibility (1 channel, 1 data window, with the sampling rate and number of samples per cycle still variable).

For the minimum versatility case the maximum output sampling rate was measured to be 250 kHz.

The maximum conversion rates for the D/A converter programmes for versatility are shown in Table 1.

The PC is not dedicated to the D/A conversion once the process is triggered, as the transputer on the D/A converter works independently of the PC. The PC can

Table I : Maximum D/A conversion rate vs number of output channels.

No of channels	1	2	3	4	5	6	7	8
Sampling rate (kHz)	81	69	59	54	48	42	38	33

thus be used for other tasks such as the interpretation of test results.

3.3. Amplifiers

A set of amplifiers was developed for the purpose of linear amplification of transient signals, which makes them particularly useful for relay-testing. To obtain the desired current for this purpose, 250 VA transconductance amplifiers (50 A_{peak} maximum) are used. These, together with 200 VA voltage amplifiers (200 V_{peak} max) provides adequate test signals to the relays. These amplifiers are fully compatible to the D/A facility described in this paper.

3.4. Analog to digital conversion.

In order to close the loop of the system as described in Figure 1, analog signals from the amplifiers and the tested relay have to be captured and interpreted by the personal computer. Analog to digital conversion systems are readily available and can be chosen in order to comply with the interpretation requirements.

4. SOFTWARE

4.1. Pascal front-end program.

A user-friendly menu-driven Pascal program is the interface between the DAC and the test data source. Any sequence of numbers can be converted to be usable by the DAC, although two wave and/or transient generators are done from within the main program, namely :

- EMTP - An interface with the much acclaimed Bonneville Power Administration's Electro-Magnetic Transient Program. This interface will be discussed Section 4.3.
- Wavegen - A versatile mathematical waveform generator, used to generate static

cycles of variable frequency, phase and amplitude.

Information is given in the program on the format of data files that is correctly interpreted by the DAC.

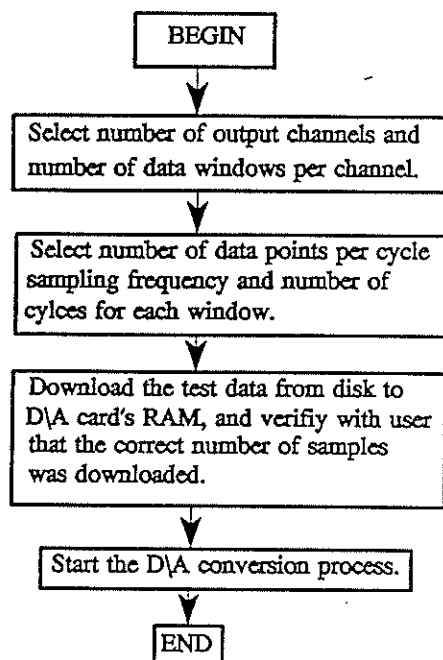


Figure 4 : Flowchart of Pascal interface options.

The Pascal routine that interacts with the OCCAM program on the DAC provides a number of parameter options as summarized in Figure 4.

4.2. The transputer code

The transputer on the D/A converter is programmed in the transputer language OCCAM, and can be changed according to different D/A converter requirements by any OCCAM programmer.

The OCCAM code consist mainly of a series of loops, synchronized by commands received from the PC (via the Link Adapter).

4.3. The EMTP interface

The Electro-Magnetic Transients Program, developed by Bonneville Power Administration, can be directly accessed from the test system. The user is shown the headings, as well as a couple of lines of simulation data from the EMTP output file, from which he can choose which current and voltage simulations he wants to use as test data. The user also has to specify which output channel to use for each set of simulated data.

The EMTP user can specify whether every simulation

point should be included in the output file, or else a skip rate is specified. The DAC-program takes this into account when the sampling frequency and number of samples per cycle is accessed. The DAC user can also specify an additional skip rate if the EMTP simulation uses a sampling frequency that is unnecessarily fast.

Certain restrictions have to be adhered to for a EMTP simulation to be used by the DAC-program, and these are stipulated before the EMTP access commences.

5. CONCLUSIONS

The programmable D/A facility described provides versatile conversion, yet leaves the personal computer free to do other tasks. The software mentioned can be replaced by more dedicated code to provide the user with the optimal system for D/A conversion.

As far as the restrictions can be adhered to, the EMTP interface provides an easy way to utilize a well-known power system simulation facility.

This system has the potential to be integrated as a versatile portable relay-testing device.

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THE DEMAND-SIDE OF AN INTEGRATED APPROACH TO ELECTRICITY PLANNING IN ESKOM

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1. INTRODUCTION

Traditionally Eskom would plan to meet a forecast demand with the most efficient combination of supply-side (generation, transmission, distribution and reticulation) resources, without making much effort to influence the demand for electricity. This strategy is generally referred to as supply-side management.

More recently, Eskom initiated an Integrated Electricity Planning (IEP) process which takes into account the supply and demand-side when planning to meet a certain forecast demand. Limaye [1] defines demand-side management as follows:

"Demand-side management involves an "intervention" by the utility on the "customer side of the meter" to influence and change the magnitude and/or shape of the customer's loads."

DSM includes covers a broad range of programs which fall into the categories of strategic load growth, load shaping and strategic conservation, explained as follows [2,3,4].

Strategic growth refers to managing the increase in electricity sales growth in a more efficient manner, e.g. encouraging a higher sales growth rate during the off-peak periods rather than during the peak periods.

Load shaping includes techniques to manage the customer's load so that the utility's overall load is less erratic and the utilisation of assets is improved. It refers to valley filling (encouraging energy usage during off-peak periods), peak clipping (reducing usage during peak times) and load shifting (maintaining the same energy consumption but at different times).

Strategic conservation strategies encourage customers to use energy more efficiently. Generally, this situation results from environmental pressures and supply capacity constraints.

This paper briefly discusses the reasons why Eskom initiated IEP, and describes the IEP process used in Eskom, with emphasis placed on the demand-side. The current and future demand-side programs are described.

2. WHY IEP IN ESKOM?

IEP has been adopted in Eskom mainly to integrate and expand supply-side and demand-side initiatives. The IEP process should ensure that supply-side and demand-side planning are performed together and that demand forecasts can be readily adapted and incorporated in the IEP as conditions change. The definition of IEP adopted by Eskom is as follows [5]:

The process which selects from a full array of demand and supply side options, that combination of actions, risks and investments which:

- *satisfies customer electricity needs*
- *is achieved at optimal value of electricity to the customer*
- *is financially viable for the utility*

The first objective ensures that the customer receives the service (s)he is prepared to pay for e.g. level of reliability, voltage quality.

The second objective ensures maximum net present value in terms of least present cost of electricity usage to all Eskom customers at acceptable risk.

The third objective ensures that the financial implications of these programs can be managed within appropriate financial constraints.

The pre-screening phase involves examining the full catalog of demand-side options to identify any specific factors or criteria that would limit the applicability of these options and to eliminate those options that are inappropriate. This screen is qualitative in nature and consists of a series of questions with weightings designed to identify measures that are applicable and possibly attractive, for example "Is the program ready for the market in terms of technology?".

In the initial screening phase, the avoided/incremental supply-side capacity and energy costs, the revenue lost/gained and the costs of researching, developing and implementing the option are used to compute the net present value of the demand-side option. The results of this initial screening process form the basis for selecting a group of demand-side options (called a program) for detailed screening as an integration of supply-side and demand-side programs.

The demand profile changes resulting from this grouping of options is used to produce the adjusted demand profile forecast which takes into account the implementation of the demand-side programs.

The adjusted demand profile is used for modeling the impact on generation, transmission and distribution costs. The least cost mix of capacity necessary to meet the forecast loads with demand-side interventions is selected.

5. DEMAND-SIDE OPTIONS FOR CONSIDERATION

Eskom is at present adopting a strategic growth strategy to address the surplus capacity situation. To encourage more efficient utilisation of the generating plant, load shaping strategies are additionally applied (e.g. time-of-use tariffs).

The strategic growth marketing drives are carried out within the market development, business development, customised pricing and electrification areas of Eskom.

Market development involves adopting marketing initiatives which influence the size and shape of Eskom's load. The market is divided into sectors (e.g. domestic, commercial, paper&pulp) and marketing

potential identified in each sector. Examples of market development programs include encouraging the use of electrode boilers, infra-red heating, heat pumps, air knives, 4 plate stoves and electric vehicles.

Business Development produces project proposals for new industrial projects and offer these to the private sector, for example the new Alusaf plant. The advantages are the establishment of new business in South Africa as well as an increase in electricity sales.

Customised Pricing sell electricity through a negotiated pricing agreement in the case where there is a win-win situation for the customer and Eskom, whilst avoiding cross-subsidisation between groups of customers. Examples include self-generation displacement agreements with various municipalities, deals with the ferro-alloy, steam and chemical industries.

Electrification refers to Eskom's program to supply electricity to all domestic customers. In the IEP context electrification is not regarded as an option. However, since its impact on the system profile is significant, IEP can be used to manage this impact.

The major thrust for load shaping emanates from the Electricity Pricing area in Eskom. The present time-of-use tariffs and the impending introduction of interruptible and real time tariffs bear evidence of this drive.

Time-of-use rates involve having different electricity prices at different times of the day to reflect the way in which the costs of production change hour by hour [7]. The time-of-use rates are based upon the time-differentiated long run marginal costs of generation. Customers take advantage of these rates by increasing electricity usage during off-peak periods and decreasing consumption during peak periods.

Interruptible rates offer to customers electricity at a reduced and specific level of reliability at a correspondingly reduced price [7]. Mitchell [8] explains that by interrupting service to selected customers for minutes or hours at a time, several times a year, the utility can reduce the amount of capacity that must be maintained for only the greatest level of demand. A potential of 3000 MW interruptible capacity in South Africa has been identified [9].

3. THE IEP PROCESS

Figure 1 illustrates the IEP process adopted by Eskom.

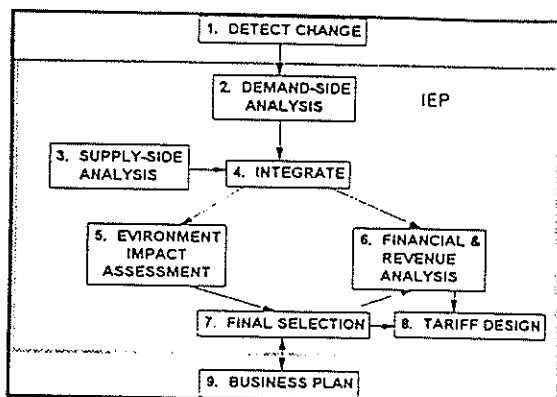


Figure 1: The IEP process overview

The demand-side and supply-side analysis involves assessing current and new options available to meet or change the forecast demand, and screening these from a technical, economic and resource availability point of view.

The integration phase involves a financial analysis of the different combinations of demand-side and supply-side options in order to select the least cost mix to Eskom. The net present value depends upon the change in generation, transmission, distribution and operations costs, the change in revenue and the costs of developing and implementing the programs.

The environment impact assessment phase examines the total environmental impact of the proposed actions, e.g. the risk, uncertainty, the impact on pollution, social and political impacts.

The financial and revenue analysis studies the effects of the changes in the supply costs and demand-side program costs on financial measures such as the pricing compact, cash flow and interest payments.

Depending upon the outcome of the environmental, financial and revenue analysis, the optimal combination of demand-side and supply-side programs is selected for implementation. The selected options are implemented, taken into account in tariff

design and the actual impact tracked and monitored and compared to the expected impact.

4. DEMAND-SIDE ANALYSIS

Figure 2 illustrates the demand-side analysis part of the IEP process.

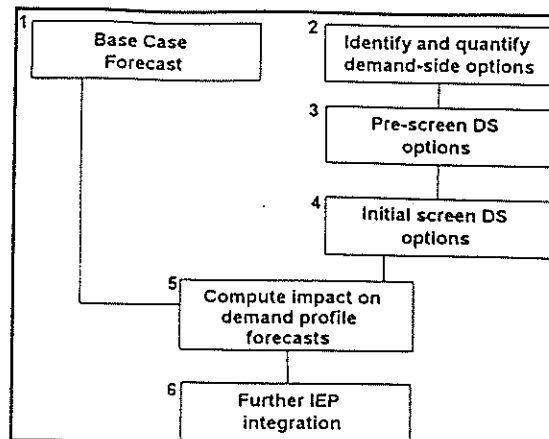


Figure 2: Demand-side planning approach

The base case forecast refers to the preparation of energy, maximum demand and hourly chronological demand profiles (excluding the influence of demand-side interventions) over a 15 year period. The hourly demand profile is required for generation expansion planning purposes. Forecasters in Eskom use econometric demand forecasting models based upon historical relationships between the system peak demand and several econometric variables; statistical trend models, and a measure of disaggregation whereby the system chronological demand profile is disaggregated into profiles for each of the major customer classes and energy forecasts are applied to the category shapes [6].

The identification and quantification of demand-side options involves listing all the demand-side options for consideration in the categories of strategic growth, load shaping and strategic conservation. Quantification involves establishing the hourly chronological demand profile of the option over its lifetime; the changes in revenue to Eskom which would result from the option being implemented; and the costs involved in researching, developing and implementing the option.

Whereas time-of-use rates typically remain fixed for at least one year, real-time pricing is based on the actual time-varying costs of electricity supply. The real time rate communicates to customers the actual cost of delivering energy as it is used [10]. Real time rates are typically announced 24 hours in advance and vary on a hourly basis.

Future load shaping programs under consideration are co-generation contracts with customers. A co-generation potential of 1500 MW in the chemical, fuel processing, paper&pulp, and metal industry has been identified [11].

With the exception of the energy efficient design awards, very little is presently being done in Eskom with respect to strategic conservation. A conservation potential of 2000 MW in peak reduction over the next 10 years has been identified [12].

6. CONCLUSIONS

IEP became real in Eskom in 1993 when the latest generation capacity expansion program took into account various demand-side measures such as interruptibility and the availability of co-generation capacity. The present demand-side emphasis is on selling more electricity to address the surplus capacity situation and load shaping to encourage the efficient utilisation of existing generation capacity. At present very little is being done on the conservation front. Due to the long time required to influence behaviour, it will be necessary to take a more pro-active stance on conservation, to ensure that the country's energy resources are used as efficiently as possible.

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A LOAD-DYNAMIC-RESPONSE BASED SAMPLING ALGORITHM FOR MEMORY LIMITED DATA LOGGING APPLICATIONS

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ABSTRACT

This paper proposes adjusting the sampling rate of a 8031-based energy-pulse-counting data-logger, according to the dynamic behaviour of the load being measured. It is proposed that this technique allows large savings in onboard memory, so that the same amount of installed RAM may be employed to capture data for longer periods of time. No loss of resolution in data is experienced with the proposed technique, although a marginally more sophisticated sampling algorithm is required. The proposed technique was implemented on a low cost hardware version of an energy measuring data logger and results obtained using this system are reported in this paper.

1 INTRODUCTION

A significant portion of performing domestic end-use device research is made up of performing physical measurements with suitable data logging equipment [2,3,6]. In stand alone data-logging applications, where fast sampling rates are required in order to measure load dynamic as well as time related behaviour, data loggers may need to be left in the field for extended periods of time. Due to the possible remote locations involved and the practical, economical, manpower and social restrictions usually incurred, as few site visits as possible are normally strived for. The total available logging time is always limited by the amount of memory installed in the logger and possible options are to install additional memory (if this is practically possible), to compromise on the time resolution of the acquired data, or then to indeed visit the site more regularly.

Once such data is down-loaded, with for example a sampling period of 1 second, the researcher may then be faced with literally thousands of identical readings, all relating no additional information. In this way, a large amount of superfluous data is generated and much valued RAM space is wasted in the process.

In order to alleviate this problem, the "jump-driven" algorithm was developed, where the sampling rate of the logger is adjusted according to ΔP (the amount

which the power being consumed has changed). This implies that during load stagnant periods (e.g. in the very early morning hours of a domestic load), the sampling rate could be reduced to one sample every couple of hours, thus saving hardware memory and preventing meaningless columns of identical values in the output files. During highly volatile periods however, the logging rate will increase, and no loss of time resolution is experienced.

2 HARDWARE DESCRIPTION

2.1 Microprocessor Section

A block diagram of the microprocessor section of the "jump driven" data-logger is presented in Fig. 1. An 8031 CPU running at 12 MHz was employed with 64 kByte of battery-backed-up static RAM serving as mass storage medium. One CPU internal 16 bit counter was employed to count energy pulses and the other was used to generate the 4800 RS232 baud-rate. The data-logger also contained a battery backed up RTC so that each of the saved samples could be date and time stamped. A 2X16 LCD display was also added as MMI to the logger. The implementation as described here had a maximum sampling rate of 1/sec and the energy pulses were accumulated for 1 second after which the counter was reset. The "jump-driven" algorithm then determined whether to log the data point or to reject it [1,5].

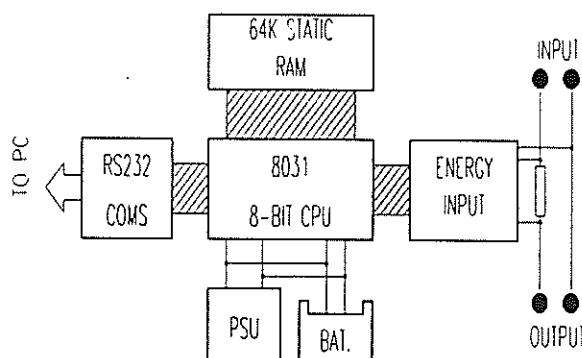


Fig. 1. Hardware layout of the microprocessor section.

2.2 Analog Input Section

The SAMES PMM101A1 power/energy meter module was employed as energy to pulse converter and this circuit calculated energy according to

$$\bar{P} = \frac{1}{t} \int_{t_1}^{t_2} ui \, dt \quad (1)$$

where $u(t)$ and $i(t)$ are the voltage and current time functions respectively. Both harmonic power and power factor were thus taken into account, and the average power integrated over approximately 1 second was output as a frequency from this module (step response of the circuit was measured as $\pm 2-3$ seconds, depending on step size). The output frequency of this module is dependant on input power, with 1 pulse being equal to 18.4 Ws. The maximum rating for the circuit employed was 1.4 kHz, which corresponded to a power rating of 25.8 kW, if pulses were counted per second. As supplied, this input module came with its own PCB and shunt and was powered directly from the 220/240VAC line. This circuit was also galvanically isolated from the CPU side by a 4N38 optocoupler.

3 SOFTWARE DESCRIPTION

3.1 The Data Logger Software

Software for the data logger was written in PLM, compiled, linked and burnt into a 27C64 EPROM. Each time the data logger saves a data point, the date and time (Day, Month, Hour, Min, Sec) for that point were also saved in BCD format. The value which was saved becomes the new threshold and as soon as the current reading differs from that amount by a pre-specified "jump value", a new data point is logged and a new threshold level is set.

Although for an 8 bit data sample application, 5 bytes need to be saved instead of one, the savings in memory remain worthwhile as was shown by the practical results (e.g. a load which stayed constant for 1 hour would require only 5 samples, where in a conventional application 3600 useless samples would have been logged).

3.2 The Personal Computer Software

Software for the PC was written in "C++" and consisted of a single input and output screen. The RTC and the "jump" value could be updated from this screen, via a RS232 link running at 4800 baud, which also served as a real time interface between the data logger and PC. All the current status bytes of the data

logger (i.e. number of samples stored, last threshold value, current measured value and jump setting) were displayed on this screen and were updated every second.

Data was downloaded from the data-logger using the same screen, into a pre-specified filename and was stored in ASCII format. The outputted data was specifically formatted for spreadsheet input and could therefore be directly imported in Excel, Lotus etc.

3.3 Post-Processing of the Data File

The combined effect of the "jump-driven" algorithm, as well as relatively slow step-response of the energy input module (s-shaped curve in Fig. 2), may lead to misleading results if data is not processed before plotting. Presentation utilities usually draw lines between sampled points and this may lead to an incorrect software plot, as is shown in Fig. 2 and Fig. 3. In order to combat these ramp effects in power levels (do not normally exist in actual loads), additional points were inserted into the data output file by a post-processing procedure. This software added additional points to the original data file and this technique is shown graphically in Fig. 2. The method therefore assumed that all variations in domestic power levels are of a step like nature. In order to further combat any incorrect results, an additional data point was also logged 1 second after all variations had stopped. This was done so that the plateau level of power, before the next jump is recorded, may be accurately established.

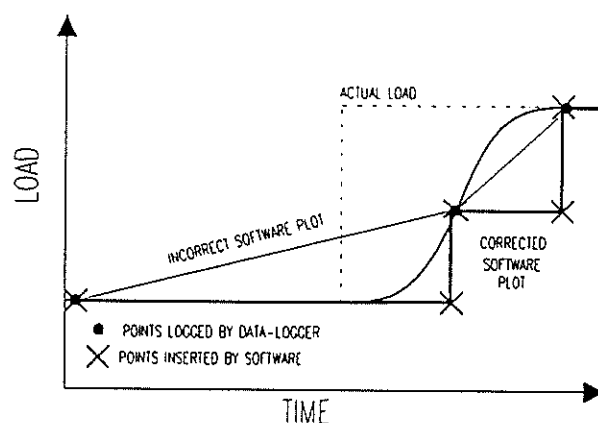


Fig. 2. Post-processing on the output data-file.

4 PRACTICAL RESULTS

In practical measurements made with the proposed system, memory savings for a five bedroom, 4 bathroom house (requiring a sampling resolution of 1 second), were in the order of 10 times, implying that

such a data logger could be left on sight 10 times as long as a conventional system with the same installed memory capacity. Data measured on this sample house between midnight and 07h30 is displayed in Fig. 3. With a resolution of 1 second, the data displayed here would normally have required 27000 samples to be processed, but the "jump driven" data logger had gathered the same information with only 600 data samples.

DOMESTIC LOAD PROFILE

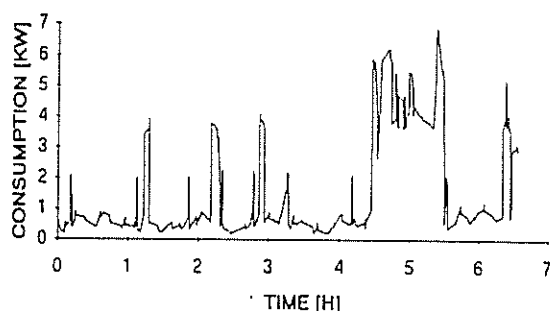


Fig. 3. The sample household between midnight and 07h00 on the 14 July 1993.

The results shown in Fig. 3. were plotted for an unprocessed output file and contain the unwanted ramp effects. The nature of the plotting algorithms used in presentation software are clearly visible and a somewhat distorted result is obtained. The load remains constant between any two sampled points and does not increase linearly as is shown in the graph. This is a direct result of the "jump-driven" algorithm and it is obvious that these ramps between individual points do not exist.

A corrected version of the retrieved data is however displayed in Fig. 4. The data displayed in Fig. 4 is an expanded version of the previous result and details electrical consumption between 3 and 4 o'clock in the morning. The jump discretization interval for these results was set at 50 W. With a higher jump level (e.g. 100 W), less data points will be sampled, but a related loss in power level resolution will then be experienced.

As the resolution of the energy input module was specified as 18.4 Ws, this was theoretically the smallest variation in input power level which could be detected. However, changes as small as these were caused by voltage variations alone, causing the "jump-driven" data logger to log continuously, thus resulting in no memory saving advantage. A compromise between accuracy and memory should therefore be established and a 100 W jump level was found to deliver satisfactory results.

DATA FOR 1 HOUR

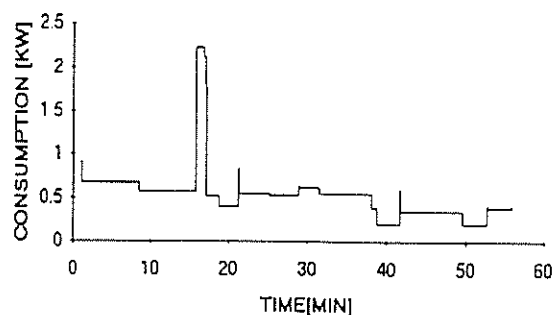


Fig. 4. Expanded view between 03h00 and 04h00 of the sample household.

5 FUTURE WORK

The hardware implementation described above consisted of discrete components, comprising the buffering latches, EPROM, RAM and the address decoding. An updated version of this data logger is however already under development and it uses a Waferscale PLD, namely the WSI 301. This single chip contains all the latches, EPROM, RAM, address decoding and two user ports in a single 44 pin PLCC package. Although external RAM will still need to be employed as mass storage, all other IC's are eliminated and the new version of the "jump driven" data-logger effectively becomes a 4 IC implementation (CPU, PLD, RTC and RS232 driver). For the updated model, a 80196 microprocessor is being used, which contains an 8 channel 10 bit A/D converter as part of the microprocessor.

A DSP power measuring algorithm is being developed for this version and it will thus avoid using the SAMES energy to pulse converter (CT's and VT's will be required instead). 8 Analog channels are used as input, with one being dedicated for voltage and the other 7 employed as current input channels. A plug-in analog front end is proposed, so that the analog input channels may also be employed as auxiliary channels for other applications. As the system also has several digital outputs available, it will also be employed to perform domestic control functions (via a zero voltage switching thyristor).

This new version of the data logger will also have a CAN device (single IC implementation) interfaced to it. This Controller Area Network device is intended to allow this data logger to be connected to an Area Network when implemented in the home and is the first step towards a practical investigation into an "intelligent home" [4,7].

6 CONCLUSION

In this paper, both the hardware and the software implementations of a "jump driven" data logger were presented. The aim of this data logger was to respond to changes in the measured power levels and so to avoid sampling vast amounts of data which contain no additional information. The interfacing of this system to a personal computer, the data retrieval procedure, as well as the development of the next version of this data logger was also presented. The work discussed forms part of an extensive research project regarding the development of the "intelligent home", where artificial intelligence (computer implementations) will assist both the suppliers and consumers of electricity, to be more energy efficient.

7 LIST OF SYMBOLS

A/D	Analog to Digital
BCD	Binary Coded Decimal
CAN	Controller Area Network
CT	Current Transformer
DSP	Digital Signal Processing
EPROM	Erasable Programmable Read Only Memory
IC	Integrated Circuit
LCD	Liquid Crystal Display
MMI	Man Machine Interface
PCB	Printed Circuit Board
PLD	Programmable Logic Device
RAM	Random Access Memory
RTC	Real Time Clock
VT	Voltage Transformer

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A FOUR LEVEL HIERARCHY OF CONTROL SYSTEMS IN AN ELECTRICITY UTILITY: IMPLEMENTATION ISSUES AND EXPERIENCES

G van Welie T N Carolin

ABSTRACT

Eskom has established a technological vision in the area of network control, called the "Four Level Control Vision". This consists of a hierarchical network of control systems, each interconnected to the other, for the control of the Transmission, Distribution and Reticulation networks and the local control of substations. The concept was first realised in 1987. It has taken six years and an estimated 300 manyears of effort by Eskom and suppliers to date to implement the concept. Since 1987, the following projects have been implemented: ENCOR (Transmission EMS), SMART (Distribution EMS/SCADA), ERTU (Substation Control), and PMRTU (Reticulation Control). This paper discusses each project briefly and elaborates on implementation issues, risk areas and learning points associated with each project.

1. INTRODUCTION

Eskom's views its power grid as subdivided into three hierarchically interconnected networks referred to as: Transmission (132kV to 765kV), Distribution (33kV to 132kV), and Reticulation (below 33kV)

The status quo by 1987 was that the Transmission network had been controlled via an EMS (Energy Management System) since the mid 1970's, whilst part of the Distribution network had been controlled via SCADA (Supervisory Control and Data Acquisition) systems since the early 1980's. Control of the reticulation network was virtually non-existent.

As a result of a process of establishing user requirements for substation control during 1988, it was realised that a broad based strategy for the control of Eskom's networks and substations was

required. This strategy was embodied in a concept named the "Four Level Control Vision". The concept of a corporate vision gave focus and direction to Eskom's efforts in the field of network and substation control and allowed suppliers to understand Eskom's intentions in this area. Integral to the "Four Level Control Vision" was the concept of an Eskom standard telecontrol protocol based on IEC standards. This protocol, called ESTEL, was to become the "glue" which cemented the elements of the technical vision together. Work was started on ESTEL in 1989 and it was implemented as a product for the first time in 1992.

2. THE FOUR LEVEL CONTROL VISION

As the name implies, the Four Level Control Vision consists of four levels in a hierarchy of control systems, each interacting with the others to facilitate the control of the Eskom power network. Figure 1 shows the levels and the present as well as proposed interconnections between levels. The levels in the hierarchy together with the systems that have been implemented to realise the vision are discussed below.

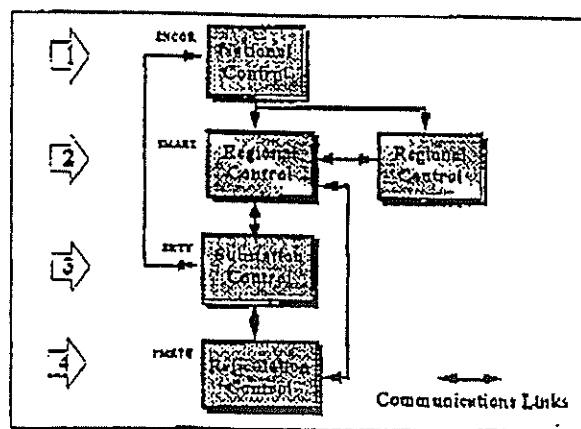


Figure 1 : Four Level Control Vision

Level 1 is the control centre for the control of the Main Transmission System, which includes the transmission network and the control of the power stations. The focus of the EMS system, named ENCOR (Eskom National COmputer Replacement), is on economic dispatch and the secure control of the power network.

Level 2 contains the control centres for the control of the distribution network, and where appropriate, the reticulation network. The project was given the acronym SMART (Standard MAster for Regional Telecontrol).

Level 3 refers to the level of substation control. Substation control in this instance can range from a simple RTU (Remote Terminal Unit) to a sophisticated ICS (Integrated Control System) which is coordinated with the new generation numerical protection relaying schemes. This project was given the acronym ERTU (Enhanced Remote Terminal Unit).

Level 4 refers to the standalone control centres and the RTUs required for the control of the reticulation network. A product was developed by two local companies to address this market. The project is referred to as PMRTU (Pole Mounted RTU).

3. LEVEL 1: ENCOR

This is the most sophisticated of Eskom's control centres and it has a team of some thirty people involved in the direct support of the system. The previous Leeds & Northrup system was replaced in 1991 with the first phase of a new EMS procured from Westinghouse U.K. The EMS functionality is based on software which Westinghouse sources from ESCA Corp. in the USA. The system is based on a modern, distributed network of DEC computers, interconnected using an Ethernet Local Area Network (LAN). The man machine interface employs the latest Graphic User Interface (GUI) technology and an upgrade to full graphics functionality is scheduled for 1993. ENCOR currently interfaces to some 130 RTUs (Remote Terminal Units), which will grow by approximately 5 RTUs per annum.

ESTEL was implemented on ENCOR for the first time in 1992. An upgrade to the communications capacity of ENCOR to cater for a new network of ERTUs is planned for 1994.

At the time of the placement of the contract (1986), a firm requirement had been the provision of full graphics software for the GUI. Full graphics software consists of pixel driven, vector based graphics, which allows features such as continuous

panning and zooming, decluttering, etc. This was very much a state-of-the-art requirement, which suppliers were claiming they could meet.

Unfortunately, no supplier in the marketplace could in reality provide this functionality at the time, and it was only in 1991/2 that stable products started appearing in the marketplace. One of the biggest risks in procuring a sophisticated EMS is the tendency to over-specify functionality requirements on the part of the electricity utility and the propensity to over-promise on the part of the suppliers. Unfortunately, due to competitive pressures in the marketplace, suppliers are under pressure to market products which do not fully exist, or are not yet fully stable. This is lightheartedly referred to as "vapourware" in the industry. A good solution to this problem is to conduct an audit of the supplier's product at the supplier's works, and to be extremely careful on relying on the so-called "demo" version of the software.

5. LEVEL 2: SMART

The contract for the provision of SMART systems was placed with Westinghouse UK in 1992. SMART systems are for all practical purposes clones of the ENCOR system. The main difference is the use of the latest DEC hardware and the provision of EMS functionality as an upgrade path, rather than as part of the core functionality. The Factory Acceptance Test (FAT) for the first SMART system for the Johannesburg Distributor is planned for December 1993. Three other distributors will install systems during 1994/5.

A problem with the acquisition of systems like ENCOR and SMART relates to the choice of hardware platform. Users of EMS systems have to accept that although the latest technology is ordered at the time of contract placement, the computer hardware will in all likelihood be obsolete by the time the system is delivered. An extremely important consideration is thus the provision of a "seamless" upgrade path by the supplier from one version of hardware to another.

A term much used in the industry today is that of the "open" system, and this has become an issue which has received a lot of attention within Eskom during the SMART project. In concept, the open system provides the user with an ideal solution: software applications which are independent of hardware platforms and operating systems, seamless interconnection with other utility information systems, fully scaleable architectures, and so forth. Unfortunately, the reality is that although great strides have been taken by suppliers in this area, there is no such thing as a truly "open" system. The

degree of "openness" will vary from supplier to supplier, and claims of "openness" should be viewed, initially at least, with circumspection. The closest one can get to independence of hardware platforms is to purchase a system which is based on the UNIX operating system. However, due to the plethora of versions of UNIX, which tend to be specific to hardware manufacturers, this does not help much. The complexity of EMS software is such that it is unlikely one would find an application which can be ported across to any hardware platform supporting "UNIX".

Practically, the next best bet is to specify that the vendor's application should adhere to a standard operating system interface such as the POSIX standard. This standard specifies the Application Program Interface (API) for calls to the operating system and the intention is that as long as vendors have complied to this interface, the application will be independent of the operating system used. The problem is that it is nearly impossible to audit a supplier's claim of POSIX compliance and as with all standards, there are obviously areas within the standard which are open to interpretation or variances in implementation by the various suppliers.

Another area of risk with regard to adhering to project schedules, is the tendency on the part of the user to completely under-estimate the effort required to configure the database for the system. This is difficult enough in the case of a SCADA system, but if EMS functionality is required, a topological database will be required which significantly increases the effort needed. Fortunately for suppliers, the configuration of the database is often the responsibility of the utility, which limits the suppliers liability in this regard.

6. LEVEL 3: ERTU

The ERTU project provides for a range of substation control products. The contract for the provision of ERTU equipment was placed with a local company, IST, during 1991. IST had tendered a system based on a Westinghouse RTU, the TALUS. However, development was required in order to meet Eskom's requirements. The ERTU project can essentially be viewed as two projects.

The first deliverable was the provision of an RTU. Development was required in the area of protocol emulation (ESTEL and PUTU), and in the area of plant interfacing. The first RTUs were installed in Eskom in August 1993.

The second deliverable is the provision of an Integrated Control System (ICS), which is

coordinated with new generation numerical protection schemes (referred to as Phase III protection). This is a leading edge application by utility standards. The first ICS system will be commissioned early in 1994, and the first Phase III protection scheme will be completed by mid 1994. One of the biggest problems in the area of ICS is the relative lack of maturity of the application, resulting in most vendor offerings being proprietary and having a distinct lack of "openness". Eskom has invested significant effort in the application and has tried, as far as practical, to steer the development to adhere to industry standards, both in terms of software and hardware.

The ICS consists of a centralised, dual redundant man machine interface and substation server configuration, interfacing to the ERTU front end system via an Ethernet LAN. Each ERTU front end processor controls up to 20 bay processors, which are essentially specially configured RTUs, located in the interface panels of the Phase III protection scheme. Each bay processor communicates via a fibre optic serial connection to the ERTU front end processor. Access to the ICS is controlled via a "gateway" ERTU which is configured as a communications processor on the LAN. The system is graphically depicted in Figure 2. below:

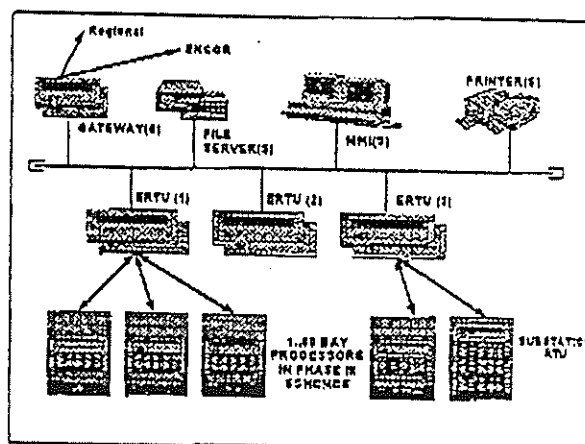


Figure 2: Integrated Control System

There are several areas in the ERTU development which required special attention;

- The repackaging of the standard RTUs to meet the fast transient burst test and voltage withstand capability specified by IEC255-4
- The complexity of the Man Machine Interface (MMI) development and application engineering. The development is based on a Microsoft Windows environment, which offers an upgrade path into the Windows NT operating system, should it be required. In addition, this provides Eskom flexibility in the choice of the hardware platform, as Windows NT will be supported by a range of hardware vendors.

- The complexity and performance of the alarm and event handling sub-systems, as well as the overall complexity of system-wide communications.
- The implementation of protocol emulation for communication between the bay processor and protection relays. This item of functionality has been deferred as a result of the lack of stability and standards in this area. This is a widely recognised area of development internationally and both the IEEE and IEC are currently investing significant effort in attempting to specify standard protocols for communication with Intelligent Electronic Devices (IEDs) such as protection relays and metering devices. The interim interface between the bay processor and the protection relays will be a hard-wired interface. This has resulted in Eskom initially having to comprise functionality such as remote electronic access to protection relays.

In a project of this nature it is advisable to work very closely with the supplier, and Eskom has seconded a number of people to IST to monitor and assist with the development. Another technique used to minimise implementation risk was the concept of rapid prototyping. This was utilised for the specification of the MMI functionality, the EMI protection of the bay processor, and the physical interfacing of the bay processor into the protection scheme interface panel.

7. LEVEL 4: PMRTU

The Pole Mounted RTU (PMRTU) system consists of three main elements, viz. a central controller, a number of PMRTUs mounted on reticulation poles in the field and a UHF radio based communication network. The PMRTU acts as an interface to pole-mounted mechanical devices (gas operated actuators).

This product was developed by two local suppliers, viz. Conlog and Spescom. The first prototype systems were delivered early in 1992, and many systems have successfully been installed since then. Eskom has recently undertaken a second phase of development, whereby the communications protocol implemented in the RTUs has been standardised. This allows RTUs from one supplier to communicate to the central controller of another supplier. The protocol utilised is a report-by-exception protocol which is optimised for the UHF radio communications medium. The RTU reports a change to the central controller as soon as a change in status has occurred. The same channel is shared by many RTUs and access to the channel is controlled via a collision sensing mechanism with a random back-off period before retry. The principle

employed can be likened to a "low speed Ethernet". The complexity of achieving a workable standardised solution was initially underestimated, however the standard protocol was finally implemented in the third quarter of 1993.

There were three main implementation issues with the PMRTU project. The first related to the packaging of the RTU to meet the IEC 255-4 Class IV EMI specifications. The second issue centred around the complete packaging of the RTU, the gas bottle, the mechanical actuator, the radio and the solar panel. A neat, workable package which was easy to commission had to be designed. The third issue related to the complexity associated with designing a report-by-exception protocol which was not based on a polled system. The principles employed have been explained above.

9. THE ROAD AHEAD

Eskom has made significant progress towards final implementation of the "Four Level Control Vision". The outstanding element is the completion of the ICS development. The focus in this area of technology is now moving from the provision and design of solutions and products to the application of these products in the field on a wide scale.

Continual adjustment and evolution of interfaces will be required to achieve the final integration of all four levels. Presently there is no linkage between Level 1 and Level 2 and between Level 2 and Level 4. The integration of PMRTU into the SMART system is currently underway, with first implementation scheduled for 1994.

Eskom has also committed to a large scale upgrade of its communications infrastructure over the next five years, which is of tremendous significance to the continued viability of the "Four Level Control Vision". Good communications capability is a precondition to the successful integration and implementation of network control systems.

10. CONCLUSION

Eskom recognised the need for a coordinated, corporate-wide approach to the implementation of network and substation control. The establishment of a technical vision gave the organisation the necessary focus in terms of its design and procurement efforts. In addition, one of the most critical aspects to the success of the whole project has been the ability to sustain continuity of involvement of key Eskom resources for the duration of the project.

User Experience with EMS-Functions

by
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ECC Europe / Switzerland

This paper reports the results from the evaluation of a CIGRE-questionnaire on "user experience with EMS-functions" which was issued in January 1991 and answered by 102 utilities all over the world. This evaluation showed that application functions are much more used and accepted today than some years ago. Nevertheless there seems to be a large area of necessary improvements. Especially the handling of the different functions and the presentation of the results of the function to the operator might be the area of future development.

1. Introduction

Energy Management Systems (EMS) have been installed in System Control Centres since the early 1970's. Most of them are equipped with Security related Energy Management Functions, such as State estimation, Contingency Analysis, On-line Load Flow and others. It is in the aim of the CIGRE Working Group 39.02 to survey the general performance of the control centres. For this purpose an earlier questionnaire has been circulated in 1985 which was answered by 98 utilities, representing 20 countries. The results of this questionnaire was published in [1].

The objectives of this first questionnaire were to determine

- the usefulness or effectiveness of EMS
- what functions or new facilities are needed for future EMS.

One of the results of this first questionnaire was, that in more than 50% of the replies listed under "Future requirements on EMS" the utilities marked the line

- must be more user oriented.

It was the objective of a second questionnaire to find out in more detail what is meant by this requirement. This second questionnaire was

102 utilities. The geographic distribution of these utilities is as follows:

Africa	3
Asia	16
Australia	6
Europe	39
North America	32
South America	6

Grouping of the utilities according to the amount of installed capacity yields the following distribution

greater than 20.000 MW	13
10.000 ... 20.000 MW	15
3.000 ... 10.000 MW	44
1.000 ... 3.000 MW	22
less than 1.000 MW	8

The questionnaire was divided in three main parts: evaluation of the "previous" control system, evaluation of the "existing" control system and future requirements. The "previous" control system was the one which was already replaced by the "existing" control system. Due to that it was hoped to find out the improvements made from one generation of control systems to the next. The evaluation follows in structure the questionnaire.

2. Evaluation of the "Previous" Control System

48 utilities had a 'previous' control system, i.e. they have already replaced an existing energy management system. The 'improvement of operational properties' were the main criteria for the selection of this previous system, followed by 'technical features' and 'performance level'. 'Commercial conditions', 'upgradeability', 'more user orientation', 'technology standard of software', 'similar installation already in commission' and 'minimum "special to project" development

required' were the requirements, which were less considered. Those systems were not accepted too well, since the mean value for the grading for the question: 'Would you consider your previous control system as "user oriented"' was on a scale from -5 to +5: 0,89 with a standard deviation of 2,9 (i.e. 50% of the answers are between -2.0 and +3.8).

The main reasons for not sufficient user orientation is the area of human-machine interface (no graphic HMI, HMI response too slow), followed by problems with the hardware (too slow, CPU too small) and with the database (poor database maintenance tools, difficult to update database, data entry and change too slow).

Most of those systems had to be replaced because it was not possible to expand either hardware, software functionality, size of the database and/or lack of horsepower of the used CPU's. It seems that there have not been many problems with the application software (or have the functions not been used or not been needed?), because the reason for replacement: 'algorithm inadequate or not trusted' got the lowest grade, i.e. minimum importance for upgrading the control system.

3. Evaluation of the "Existing" Control System

The "Existing Control System" was evaluated in the following way: the SCADA-functions was evaluated with highest of satisfaction, followed immediately by the application functions. The area of human-machine-interface (lack of engineering tools) got the lowest grade of satisfactions as well as the flexibility and maintainability of the used database and the expendability of the hardware.

The functions installed vary considerably with respect to the different geographical areas. In Asia and North America the functions which are installed in more than 75% of the utilities are Economic Dispatch, Reserve Monitoring and the SCADA-functions (in North America also Interchange Transaction Scheduling). In Europe the situation is quite different: SCADA functions are installed in 96% of the utilities followed immediately by Topology, Operator

Load Flow and State Estimation. It seems that in Asia, in particular in Japan, because most of the answers from Asia are from Japan, and in North America much more emphasis is placed on the so called economic functions. In Europe power system monitoring is more important. Operator Load Flow and State Estimation are the functions mostly used. Economic Dispatch and Production Costing for instance are installed in 91 resp. 76% of the utilities in North America, while in Europe only in 32% of EMS have these functions included.

On the other hand Dispatcher Training Simulator, Optimal Power Flow, Modelling of the External Network, Load Forecast and Thermal Unit Commitment are much more frequently installed in North America than in Europe.

The evaluation of the existing Application Functions yields a large acceptance and satisfaction with State Estimation and Operator Load Flow. Not only consistency, quality and usefulness of the results of these two function got very good grades, also the presentation of the results is graded relatively high. This is not the case for Contingency Analysis. The usefulness and the presentation of the results of Contingency Analysis have to be improved in the future considerably. Today the grades are rather low. The lowest grade got the 'on-line help' functions (-1 on a scale from -5 to +5), which means that most of the utilities are not satisfied with those functions or they are not available at all.

Most of the utilities (65%) have modified the application software delivered by the vendor of the energy management system, because their needs have changed. Only 15% did not modify anything. 71% of the utilities have also developed new application functions which were not delivered by the vendor. 20% did not develop anything by themselves. 5 utilities stated that they had developed the whole control system 'in house' without any vendor.

It seems that the question "which quantifiable benefits have you realised from the application software" was rather difficult to answer. 14% said that the operating cost had been reduced, 14% stated that transmission losses could be kept at a minimum and 9% admit that the benefits are not quantifiable at all. Other benefits, which were mentioned are:

- better operating conditions, especially in difficult situations
- more security (whatever it means)
- better monitoring of power system state
- better modifications
- better restoration (mainly due to Dispatcher Training Simulator)
- better data consistency
- higher loading of the power system,
- better voltage profile.

A very large variation - from 1 man-year up to 100 man-year showed the evaluation of the manpower required from the utilities for the installation of their system. For EMS made by vendors the average manpower required from the utilities is 12 man-years. EMS made by the utilities themselves resulted in an average manpower of 100 man-years. Those figures depend very much on the size and the required functions of a specific utility.

The time required for installation of systems made by vendors averages to 2,4 years, systems made by the utilities needs in the average 6 years.

The manpower required to maintain the application software is in the average 4 man-year per year. Here it is a large difference between Asia and the rest of the world. In Asia the average manpower required is only 1 man-year per year.

Large utilities need much more manpower for the maintenance than small utilities (ratio 7 : 1).

In most of the utilities the application programs are an integral part of system operation (68%) and 'daily operation needs these programs'. Only few answers to that question are negative and the reason stated is again in most of the cases a poor human-machine interface.

The amount of training necessary for each user for the application functions is in average 4 man-weeks (In Europe in average 6 man-weeks).

It is interesting to notice, that the "existing" EMS have significantly improved and are considered to be much more user oriented than the "previous" ones. The average value for grading user orientation of the existing EMS on a scale from -5 to +5 is 2,6 with a

standard deviation of 2,3 (i.e. 50% of the answers are between 0,3 and 4,9) compared to 0,89 and 2,9 respectively for the "previous" generation of control systems. Easy to use human-machine interface and fast response are the main reasons to consider an EMS as "user oriented".

The EMS-functions which are mostly used by the operator as expected are the SCADA functions, Reserve Monitoring, Interchange Transaction Scheduling and Energy Accounting. State Estimation and Operator Load Flow seem to be used in Europe more than in North America, while Interchange Transaction Scheduling integrated into an EMS is used much more in North America than in Europe. The function which seems to cause a lot of problems all over the world is Optimal Power Flow. Most of the utilities claim, that it is not used by the operator as expected.

The most frequent reason for not using regularly a certain function, is "results not so often needed by the operator". This means, that the particular function itself is not needed by the operator.

For the most important functions the reasons for not using them regularly are:

- State Estimation:
 - results not so often needed by the operator 15%
 - not enough data available from the on-line system 8%
- Operator Load Flow:
 - results not so often needed by the operator 16%
 - not enough data available from the on-line system 12%
- Modelling of the External Network:
 - results not so often needed by the operator 17%
 - not enough data available from the on-line system 12%
- Network Reduction:
 - results not so often needed by the operator 35%
- Contingency Analysis:
 - results not so often needed by the operator 25%

- not enough data available from the on-line system 10%
- on-line Short Circuit Calculation:
 - results not so often needed by the operator 35%
- Optimal Power Flow:
 - MMI too complicated, needs too much interactive operation 23%
 - algorithm not working satisfactorily 19%
 - results not so often needed by the operator 15%
- Voltage-VAr Scheduling:
 - algorithm not working satisfactorily 14%
 - results not so often needed by the operator 14%
 - not enough data available from on-line system 11%
- Load Forecast:
 - results are not correct 11%
- Thermal Unit Commitment:
 - MMI too complicated, needs too much interactive operation 12%

If all reasons for not using a particular function as expected are summed up, the following functions top the list:

- Optimal Power Flow
- Voltage-VAr Scheduling
- Network Reduction
- Contingency Analysis.

The departments within a utility which are also using the results of application functions of an EMS are:

- Billing:
 - Energy Accounting 54%
- Planning:
 - Hydro-Thermal Co-ordination 42%
 - On-line Short Circuit Calc. 39%
 - Fuel Resource Scheduling 35%
 - Operator Load Flow 34%
 - Network Reduction 29%
 - Thermal Unit Commitment 27%
 - Load Forecast 26%
 - Contingency Analysis 25%
 - State Estimation 25%
- Management:
 - Fuel Resource Scheduling 24%

- Maintenance:
 - Autom. System Trouble Analy. 21%

The main criteria for selection of a control system from a certain vendor were (graded from -5 to +5):

- technical features 2,8
- improvement of operational properties 2,6
- vendor's reputation and experience 2,4
- performance level 2,4
- maintainability 2,2
- technology standard of hardware 2,0
- more user oriented 1,8
- upgradability 1,8

The lowest rating of importance in that context got again (as for the 'previous' control system) the criteria:

- minimum "special to project" development required
- similar installation already in commission
- delivery schedule.

More than 50% of the EMS contain training facilities in one way or another, 36% have a Dispatcher Training Simulator, most of them with dynamic models of the elements of the power system.

The problems encountered during implementation and operation of the different functions can be grouped in problems with

- data
- algorithm
- human-machine interface
- documentation
- training.

The problems encountered during implementation for the most important functions have been:

- State Estimation:
 - not enough data from own system 26%
 - data not accurate enough 25%
 - not enough data from neighbouring system 15%
- Operator Load Flow:
 - not enough data from own system 15%

- Modelling of the External Network
 - not enough data from neighbouring system 34%
- Optimal Power Flow:
 - algorithm not robust enough 19%
 - results not accurate enough 15%
 - presentation of results not sufficiently user oriented 15%

The problems encountered with the different function during operation are sometimes quite different:

- State Estimation:
 - not enough data from neighbouring system 12%
- Modelling of the External Network:
 - not enough data from neighbouring system 25%
 - user documentation inadequate 14%
- Contingency Analysis:
 - presentation of results not sufficiently user oriented 16%
- Optimal Power Flow:
 - execution time too long 12%
 - algorithm not robust enough 12%
 - presentation of the results not sufficiently user oriented 12%
 - operator lacking sufficient training 12%
- Load Forecast:
 - results not accurate enough 12%

It seems important to see the problems with the different functions, but much more, to see the complementary, i.e. which problems are not listed here. Most of the difficulties with implementation and operation of on-line functions are due to data problems - not enough or not accurate enough data are available to start with. This is natural and corresponds with the general opinion, that for instance State Estimation pays off half of its costs already during implementation, since it is normally the first time that the on-line data of a certain power system are checked for consistency.

Generally it can be stated, that the application functions have grown out of their infancy and most of them are used quite regularly and

successfully for the every-day work of the operator.

4. Future Requirements

It is clear that the development of the application functions is still continuing and improvements for future EMS are expected. Those required and expected improvements of the different areas are graded on a scale of importance from -5 to +5 as follows:

- alarm handling, alarm compression 2,0
- data base management system with on-line maintenance 1,8
- use of full graphic MMI 1,8
- more graphic representation 1,8
- data compression during emergencies 1,4
- use of Optimal Power Flow 0,8
- better Load Forecast 0,8
- additional functionality in case of emergency 0,8
- facilities for operator training 0,8

The specific wishes of necessary improvements of the different EMS functions are listed below. This list is ordered by the number of responses but is by far not complete, only multiple answers and interesting recommendations are mentioned:

- State Estimation:
 - automatic correction of measurements
 - identification of topological errors
 - parameter estimation
- Modelling of the External Network:
 - more real-time data from the neighbouring system
- Contingency Analysis:
 - more 'intelligent' application required, too much hand dressing
- Optimal Power Flow:
 - clear recommendations to the operator for feasible solution
- Thermal Unit Commitment:
 - link to real-time plant availability
- Automatic Generation Control:
 - improve algorithm to better anticipate change of load

The functions which seem to have reached a sufficient maturity level (graded on a scale of importance between -5 and +5) are:

- Mimic boards and/or larger scale displays -0,8%
- SCADA functionality -0,8%
- Load Frequency control -1,2%

The additional function which are required by more than 25% of the utilities are:

- Optimal Power Flow 34%
- Dispatcher Training Simulator 33%
- Alarm Compression 31%
- Voltage-VAr Scheduling 28%

Again there are rather large discrepancies between the different geographical areas. In Asia the most important function to be added in the future is the Optimal Power Flow, in Europe it is the Dispatcher Training Simulator followed by Optimal Power Flow and in North America it is Alarm Compression, followed by Dispatcher Training Simulator and Optimal Power Flow.

33% of the utilities want to have the possibility to connect an expert system to the EMS without specifying the problem, which should be solved by such a system. This indicates interest in new technologies.

5. Conclusion

The answers within the questionnaires showed a rather large variation which might have different reasons:

- different interpretation of the same question
- different background of the person who answered the question
- different types of utilities
- different structures of organisation of utilities.

It was necessary to formulate the questions in a negative way but the answers should be read differently. If 15% of the utilities are not satisfied with a certain function, that mean that 85% of the utilities are more or less satisfied with that function. The answers showed in general a very promising trend. The development of the algorithms seems to have matured, the acceptance of those tools by the operators has been improved considerably and they are now part of the daily operation of the

power systems. On the other hand the presentation of the results and the handling of the different functions needs still considerable improvement.

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MATHEMATICAL MODEL OF THE ASYNCHRONOUS INDUCTION GENERATOR USED FOR VARIABLE SPEED BATTERY CHARGING APPLICATIONS

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ABSTRACT

The mathematical model is based on the equivalent circuit of the induction generator which in this case is further developed to compensate for varying equivalent circuit parameters. Since the machine operates under varying speed conditions, the frequency changes affects all the frequency dependent circuit parameters. The original equivalent circuit was then modified to form a new equivalent circuit whereby all the circuit parameters are referred to those at rated frequency.

1. INTRODUCTION

The mathematical model was developed as part of a wind energy project which meant that parameters which would under normal conditions be constants would now become variables. A further complication was that a standard squirrel cage induction motor was used to function as an asynchronous induction generator ie the electrical energy source of the wind-driven generator. The squirrel cage motor will therefore from here on be referred to as the generator.

The model, once constructed may be used to predict and analyse the performance of a generator under various conditions and speeds.

2. THE MODIFIED EQUIVALENT CIRCUIT

The modified equivalent are based on the standard equivalent circuit but incorporates parameters which are referred to rated frequency, in effect each condition will have its own equivalent circuit.

Let a = the ratio of generated frequency to rated frequency.

ie $a = 2\pi f_{gen} / 2\pi f_{rated}$

Let b = ratio of actual rotor speed to synchronous speed.

$$\begin{aligned} \text{ie } b &= N_{rotor} / N_{synch} \\ &= 2\pi f_{rotor} / 2\pi f_{rated} \end{aligned}$$

Therefore the slip now appears as : $s = (a-b)/a$

The equivalent circuit parameters including load and excitation capacitance referred to rated frequency thus appears as shown in figure 1.

R_m has been omitted since it has little effect and simplifies calculations.

The parameter R_r/s can be written as: $R_r/s = R_r \cdot a / (a-b)$

Omitting slip yields: $R_r/as = R_r/(a-b)$

Dividing through by " a " yields the circuit model in figure 2.

Replacing R_l/a and $-jX_c/a^2$ with a series branch simplifies the circuit further, as shown in figure 3, yielding R_E and X_E .

$$\begin{aligned} \text{Where: } R_E &= R_l \cdot X_c^2 / (a^2 \cdot (a^2 R^2 + X_c^2)) \\ \text{And: } X_E &= R_l^2 \cdot X_c / (a^2 \cdot R_l^2 + X_c^2) \end{aligned}$$

3. EQUILIBRIUM EQUATIONS

Since there is no current or voltage source in the network, the circuit must conform to the laws of conservation of real and reactive power. Hence the sum of the admittances of the three branches equates to zero:

$$Y_{stator} + Y_{magnetising} + Y_{rotor} = 0$$

$$\text{Where: } Y_s = 1 / (R_s/a + R_E + j(X_s - X_E))$$

$$\text{Therefore: } Y_s = (R_s/a + R_E - j(X_s - X_E)) / ((R_s/a + R_E)^2 + (X_s - X_E)^2) \quad \dots(1)$$

$$Y_m = -j/X_m \quad \dots(2)$$

$$Y_r = 1 / (R_r/(a-b) + jX_r)$$

$$\text{Therefore: } Y_r = (R_r / (a-b) - jX_r) / ((R_r/(a-b))^2 + X_r^2) \quad \dots(3)$$

Separating for the real and imaginary parts and equating to zero yields equilibrium equations which describe the operating point at excited frequency.

$$\begin{aligned} \text{Real:} & \quad (R_s/a + R_E)/((R_s/a + R_E)^2 + (X_s - X_E)^2) + R_r/(a-b)/((R_r/(ab))^2 + X_r^2) = 0 \quad \dots(4) \\ \text{Imaginary:} & \quad -(X_s - X_E)/((R_s/a + R_E)^2 + (X_s - X_E)^2) - 1/X_m - X_r/(R_r/(ab))^2 + X_r^2 = 0 \quad \dots(5) \end{aligned}$$

Using the real part of the equation and substituting in for RE and XE, the denominator cancels to zero, therefore:

$$\begin{aligned} (R_s R_l^2 a^2 + R_s X_c^2 + R_l X_c^2)(R_l^2 a^3 + a X_c^2)(R_r^2 + X_r^2(a-b)^2) + R_r^* \\ (a-b)[(R_s R_l^2 a^2 + R_s X_c^2 + R_l X_c^2)^2 + (X_s R_l^2 a^3 + X_s X_c^2 a - a X_c R_l^2)] = 0 \quad \dots(6) \end{aligned}$$

4. POLYNOMIAL IN "a"

After simplifying and arranging in descending powers of "a", a fifth order polynomial is obtained, the coefficients of which are:

$$\begin{aligned} \text{Polynomial:} & \quad Q_0, Q_1, Q_2, Q_3, Q_4, Q_5. \\ & \quad Q_5 a^5 + Q_4 a^4 + Q_3 a^3 + Q_2 a^2 + Q_1 a + Q_0 = 0 \quad \dots(7) \\ \text{Where:} & \quad Q_0 = 0 \\ & \quad Q_1 = -b R_r R_l^2 / (R_s + R_l)^2 \dots(8) \end{aligned}$$

$$Q_1 = R_r R_l^2 / (R_s + R_l)^2 + (R_s + R_l)(R_r/R_l)^2 + b^2 (R_s + R_l)(X_r/R_l)^2 \quad \dots(9)$$

$$Q_2 = 2b(R_s + R_l)(X_r/R_l)^2 - b R_r [(R_s/X_c)^2 + (X_s/R_l)^2 - 2(X_s/X_c)] \quad \dots(10)$$

$$Q_3 = R_r [(X_s/R_l)^2 + (R_s/X_c)^2 - 2(X_s/X_c)] + (R_s + R_l)(X_r/R_l)^2 + R_s(R_r/X_c)^2 + b^2 R_s(X_r/X_c) \quad \dots(11)$$

$$Q_5 = R_r(X_s/X_c)^2 + R_s(X_r/X_c)^2 \quad \dots(12)$$

$$Q_4 = -b(Q_5 + R_s(X_r/X_c)^2) \quad \dots(13)$$

The positive real root of this equation yields the value of "a" for the given circuit parameters. From this value of "a", the stator frequency under these conditions may be calculated.

5. MAGNETISING REACTANCE

Solving for X_m in (5) and substituting into (4), an expression to find the magnetising reactance X_m at rated frequency is obtained.

$$X_m = (R_E + R_s/a) * [X_r^2 + (R_r/(a-b))^2] / [(R_r/(a-b))(X_s - X_E)X_r * (R_E + R_s/a)] \quad \dots(14)$$

6. NO LOAD CONDITIONS

On no load conditions $R_l = \text{infinity}$, $R_E = 0$ and $X_E = X_c/a^2$. Under these conditions the polynomial no

load coefficients are:

$$Q_{00} = -b R_r \quad \dots(15)$$

$$Q_{01} = R_r \quad \dots(16)$$

$$Q_{02} = -b R_r [(R_s/X_c)^2 - 2(X_s/X_c)] \quad \dots(17)$$

$$Q_{03} = R_r [(R_s/X_c)^2 - 2(X_s/X_c)] + R_s(R_r/X_c)^2 + b^2 R_s(X_r/X_c)^2$$

$$Q_{05} = R_r(X_s/X_c)^2 + R_s(X_r/X_c)^2 \quad \dots(18)$$

$$Q_{04} = -b[Q_{05} + R_s(X_r/X_c)^2] \quad \dots(19)$$

From (20) let $R_E = 0$ and $X_E = X_c/a^2$ to get an expression for X_m under no load conditions:

$$X_{om} = (R_s/a) * [X_r^2 + (R_r/(a-b))^2] / [(R_r/(a-b)) * (X_s - X_c/a^2) - R_s R_r/a] \quad \dots(20)$$

7. OUTPUT VOLTAGE

From the equivalent circuit, in figure 2, it can be seen that the output voltage can be obtained from:

$$V/a = E_g / (R_E + R_s/a + j(X_s - X_E)) * (R_E + jX_E)$$

The absolute value of which is:

$$V = a * E_g * [(R_E^2 + X_E^2) / ((R_E + R_s/a)^2 + (X_s - X_E)^2)] \quad \dots(21)$$

After substituting in for RE and XE and rearranging, the terminal voltage per phase can be obtained from the expression:

$$V = a * E_g / [(1 - a^2(X_s/X_c))^2 + a^2((X_s/R_l)^2 + (R_s/X_c)^2 + (1 + (R_s/R_l))^2 - 1)^{0.5}] \quad \dots(22)$$

8. NO-LOAD OUTPUT VOLTAGE

$$V = a * E_g / [(1 - a^2(X_s/X_c))^2 + (a * R_s/X_c)^2]^{0.5} \quad \dots(23)$$

9. CONCLUSIONS

The mathematical model was found to be a useful tool for determining the values of peripheral components to match a squirrel cage induction motor characteristics, when used as an asynchronous induction generator, to that of a wind turbine once certain basic parameters are obtained. Its authenticity and accuracy is based on actual motor/generator parameters obtained by extensive test bench procedures.

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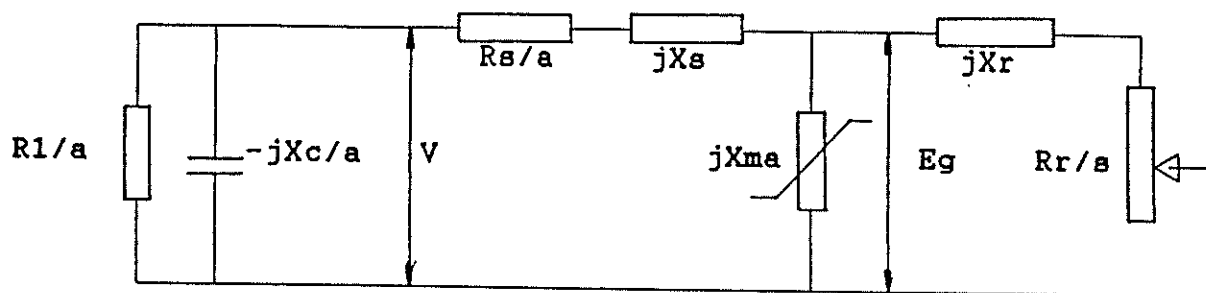


Figure 1.

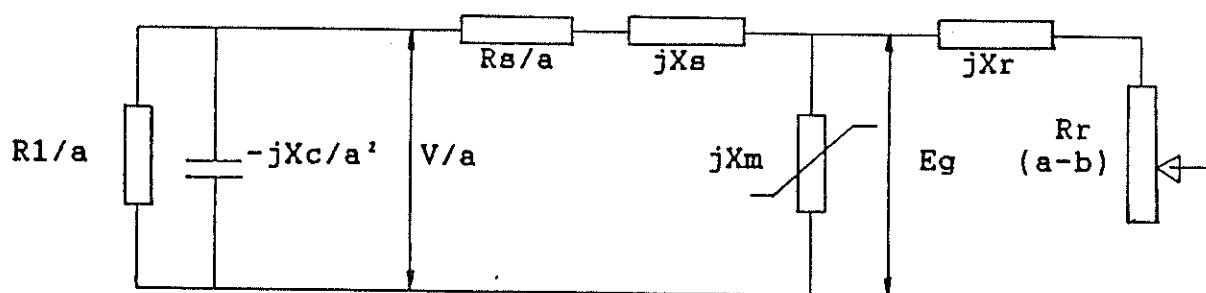


Figure 2.

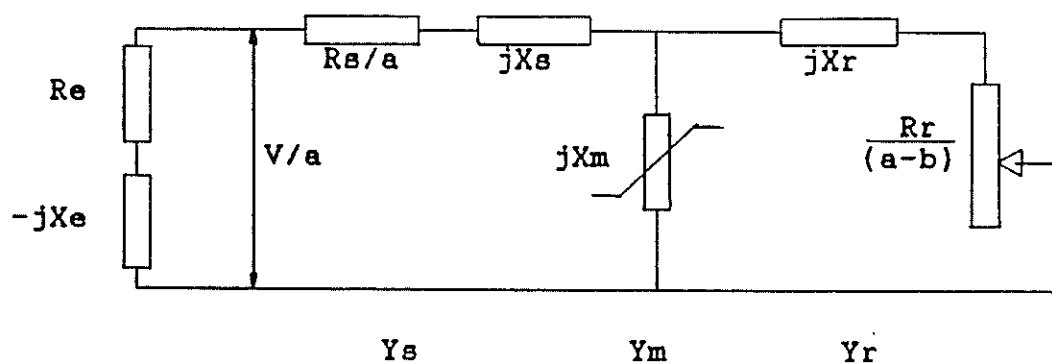


Figure 3.

A COMPARISON OF THE EFFICIENCY OF A.C. MOTORS: THE CAGE INDUCTION, THE SOLID-ROTOR INDUCTION, AND THE RELUCTANCE SYNCHRONOUS MOTOR

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Abstract

The aim of the paper is a comparative analysis of three different types of similarly rated a.c. motors. The analysis was done on the basis of efficiency of a.c. motors with unexcited rotors. The three motors tested were the cage induction, the solid-rotor induction and the reluctance synchronous motor. The same stator of a 4 pole 1.5 kW induction motor has been used. The effect on efficiency from driving the motors from a sinusoidal or inverter fed supply have also been investigated.

It has been shown that the cage induction motor has the best efficiency and that other simplified rotor constructions do not improve the efficiency.

1 Introduction

The comparative analysis of a.c. motors with unexcited rotors is aimed at construction of a simple and cheap a.c. motor for mass applications. At present time, the cheapest and the most frequently used electric motors are cage induction motors. The replacement of the cage rotor with a cylindrical solid rotor converts a cage induction motors into a solid-rotor induction motor while the replacement of the cage rotor with a salient-pole solid rotor converts it into synchronous reluctance motor. Both rotors can be made from a piece of mild steel, not carry any windings or permanent magnets, and are cheaper than a laminated rotor with cage windings.

2 Performance of a.c. motors with unexcited rotors

2.1 Cage induction motor

The electromagnetic torque developed by the cage induction motor is proportional to the input voltage squared, i.e.

$$T_d = \frac{m_1}{2\pi n_s} \times \frac{V_1^2 (R'_2/s)}{[R_1 + (R'_2/s)(1 + \tau_1)]^2 + [X_1 + X'_2(1 + \tau_1)]^2} \quad (1)$$

where m_1 is the number of stator phases, n_s is the synchronous speed, V_1 is the input phase voltage, $R_1 + jX_1$ is the stator winding impedance, $R'_2/s + jX'_2$ is the rotor winding impedance referred to the stator system, s is the slip for fundamental harmonic, $\tau_1 \approx X_1/X_g$ is Heyland's coefficient for the stator, and X_g is the airgap (mutual) reactance.

The losses in induction motors consists of the stator winding losses ΔP_1 , rotor winding losses ΔP_2 , stator core losses ΔP_{Fe1} , rotor core losses ΔP_{Fe2} (important only in inverter-fed motors), rotational losses ΔP_{rot} , and stray losses ΔP_{str} . The stray losses according to IEC standards are estimated as 0.5 % of the output power P_{out} .

The main disadvantages of induction motors are: lower efficiency and *output power-to-volume* ratio than in synchronous motors, small airgap, possibility of cracking the rotor bars due to hot spots at heavy starting, braking, and reversal, and the necessity of the power factor correction.

2.2 Solid-rotor induction motors

When the cage rotor is replaced by a solid rotor the starting current is reduced and the rotor construction is simplified. Other advantages include: (a) possibility to obtain steady-state stability and linearity of torque-speed characteristics throughout the entire speed range; (b) high reliability; (c) high mechanical integrity, rigidity and durability; (d) low level of noise and vibrations (no slots).

On the other hand, a solid-rotor induction motor has lower output power, efficiency and power factor, and higher no-load slip than a cage induction motor of the same size. The high impedance of a solid rotor is the main reason for these disadvantages. The solid rotor impedance can be diminished in one of the following ways: (i) solid rotor made of a ferromagnetic material with the ratio of magnetic permeability to electric conductivity as small as possible; (ii) layered (sandwiched) structure of the rotor made of appropriate ferromagnetic and nonmagnetic high conductivity materials; (iii) optimum control system.

The electromagnetic torque developed by an induction motor with a solid smooth homogeneous rotor can be found by replacing R'_2/s in eqn (1) by [1]

$$\frac{R'_2(s)}{s} = a_R k_{tr} k_Z \frac{L_i}{\tau} \sqrt{\frac{\pi f \mu_o \mu_r}{\sigma}} \frac{1}{\sqrt{s}} \quad (2)$$

and by replacing X'_2 in eqn (1) by

$$\frac{X'_2(s)}{s} = a_X k_{tr} k_Z \frac{L_i}{\tau} \sqrt{\frac{\pi f \mu_o \mu_r}{\sigma}} \frac{1}{\sqrt{s}} \quad (3)$$

where $a_R \approx 1.45$, $a_X \approx 0.85$ [2], $k_{tr} = 2m_1(N_1 k_{w1})^2/p$, N_1 is the number of stator turns per phase, k_{w1} is the stator winding factor for fundamental, p is the number of pole pairs, $k_Z \approx 1 + 0.5\tau/L_i$, L_i is the effective length of stator stack, τ is the pole pitch, f is the motor input frequency, μ_o is the magnetic permeability of free space, μ_r is the relative magnetic permeability of the rotor, and σ is the electric conductivity of the solid rotor.

The losses in a solid-rotor induction motors are the same as those in a cage induction rotor. The stray losses ΔP_{str} are higher than 0.5 % P_{out} since the rotor

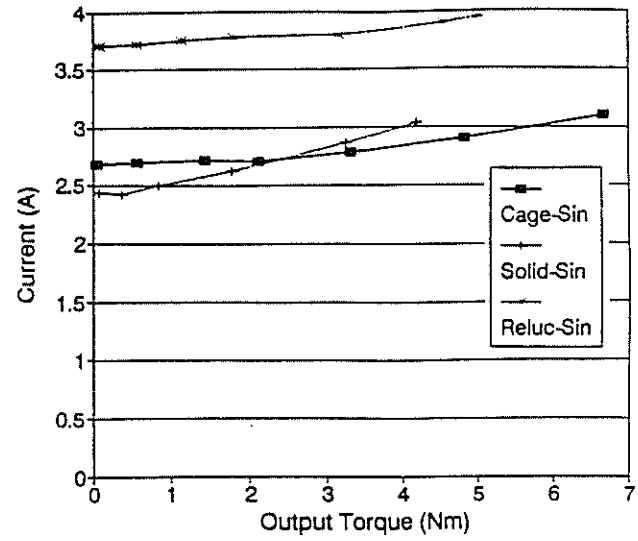


Figure 1: Comparison of the input currents for $f = 50$ Hz, $V_1 = 350$ V, and sinusoidal excitation.

surface has a higher conductivity for eddy currents induced by the higher harmonic stator fields.

2.3 Synchronous reluctance motors

The synchronous reluctance motors run with unexcited rotor having unequal reluctance in d and q axes. The electromagnetic reluctance torque

$$T_{drel} = \frac{m_1 V_1}{4\pi n_s} \left(\frac{1}{X_{s,q}} - \frac{1}{X_{s,d}} \right) \sin 2\delta \quad (4)$$

where $X_{s,d}$ is the d -axis synchronous reactance, $X_{s,q}$ is the q -axis synchronous reactance, and δ is the power angle, which is proportional to the input voltage. Small synchronous reluctance motors have a rather high stator winding resistance R_1 that is commensurable with $X_{s,d}$ and $X_{s,q}$. That is why a more precise equation than eqn (4) should be used [3].

A heavy magnetizing current resulting in high input current is due to a high reluctance of the magnetic circuit in q axis. Correspondingly, it affects the efficiency because of heavy power loss in the stator winding. The only losses in the rotor are core and stray losses.

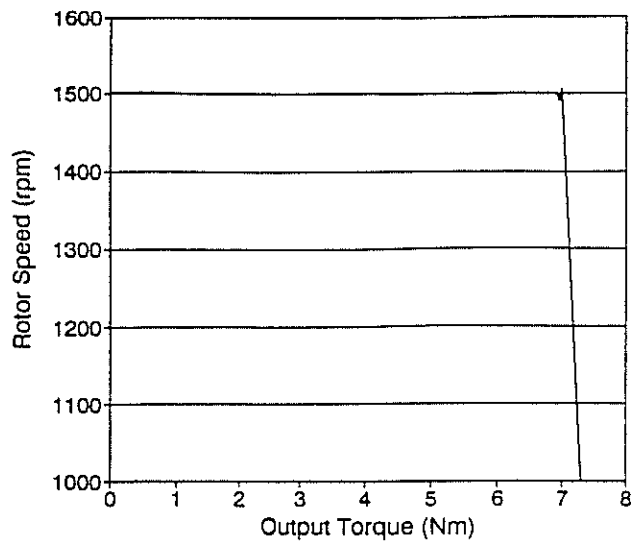


Figure 2: Speed-torque characteristic for the reluctance motor at $f = 50$ Hz, $V_1 = 350$ V, and sinusoidal excitation.

3 Comparative analysis

A small 1.5-kW, 4-pole, 1410-rpm, 380-V, three-phase cage induction motor manufactured by GEC (South Africa) has been selected for testing. According to the manufacturer, the full load torque is 10.2 Nm, the full load input current is 3.9 A, the full load efficiency is 75 %, and the full load power factor is 0.78. In addition to the commercial cage rotor, a cylindrical solid rotor and salient-pole solid rotor has been made. The airgap has been kept constant for the all three rotors ($g = 0.6$ mm). The pole-shoe width b_p of a salient 4-pole rotor 10 mm and the rotor diameter is 41 mm. The ratio of pole arc-to-pole pitch $b_p/\tau = 0.54$

Fig. 1 shows the input current measured for the three different motors with the same stator. The highest input current is for the reluctance motor and is approximately the same for cage and solid-rotor induction motor.

The maximum torque (7 Nm) of the reluctance rotor shown in Fig. 2 is lower than that of the cage

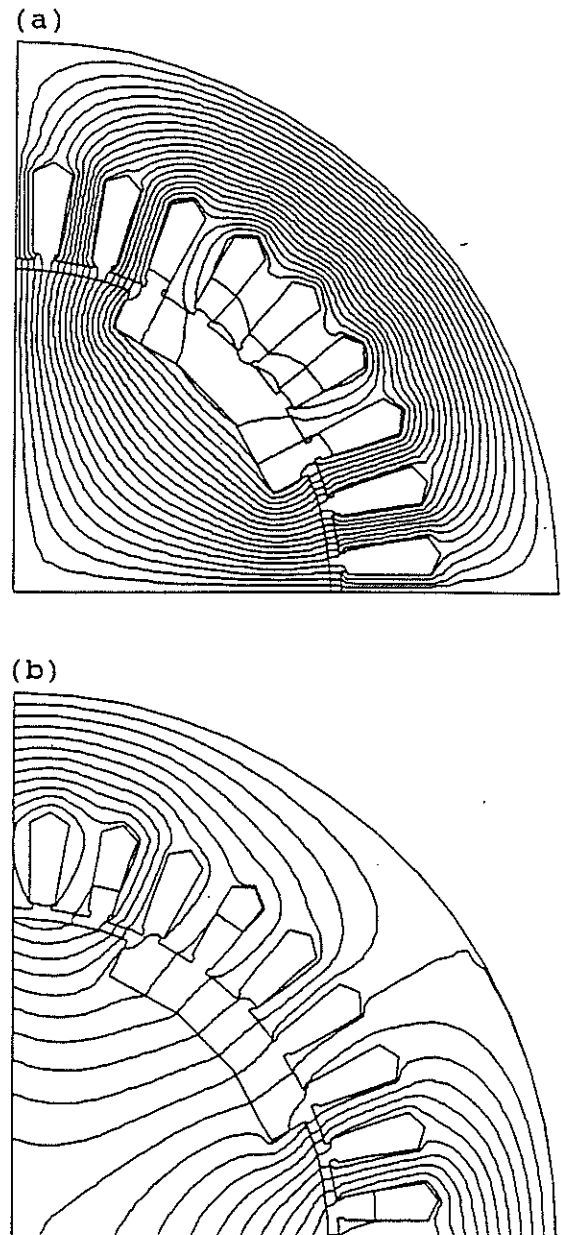


Figure 3: Magnetic flux distribution in the cross section of the reluctance motor for: (a) no-load, (b) 7 Nm ($f = 50$ Hz, $V_1 = 350$ V, sinusoidal excitation).

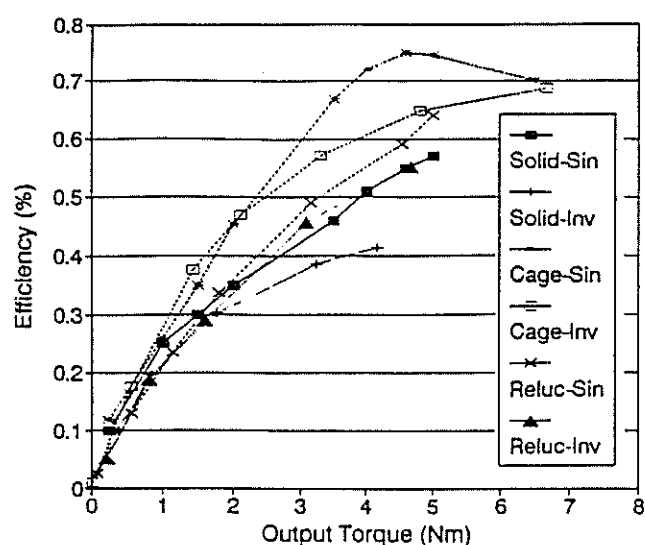


Figure 4: Efficiency-shaft torque curves for sinusoidal and inverter excitation ($f = 50$ Hz, $V_1 = 350$ V).

induction motor (10 Nm). The reluctance motor is a self-starting motor only for very low load torque (below 1 Nm) and is pulled out of synchronism if the load torque exceeds 7 Nm (stability limit). The magnetic flux distribution for no-load and maximum torque in the cross section of the reluctance motor is shown in Fig. 3. This flux distribution has been obtained with the aid of the finite element method.

The efficiency of the all three motors is shown in Fig. 4. The best efficiency is for commercial cage induction motor and the worst efficiency is for the solid rotor induction motor.

The tested motors have been also excited with a square-wave, six-pulse, voltage-source inverter. An inverter excitation reduces the efficiency due to additional higher time harmonic losses.

4 Conclusion

The attempts of simplifying the construction of a.c. motors with unexcited rotors have not been successful. It is clear that the superior rotor of the cage

induction motor means this motor out performs the solid-rotor induction and synchronous reluctance motors. The synchronous stability of the synchronous reluctance motor was also shown to be lower than the maximum torque of the cage induction motor.

The efficiencies of all three motors were reduced when the sinusoidal supply was replaced by an square wave inverter fed supply.

More research and measurements using precise instruments are required on different size motors before extensive conclusions can be drawn.

Acknowledgements

This work has been supported by the Foundation for Research Development (FRD), Eskom, and the University of Cape Research Committee.

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THE EVALUATION AND SIMULATION OF SHORT CIRCUIT CURRENTS IN INDUCTION MOTORS WITH DEEP BARS.

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ABSTRACT

Large induction motors can contribute a significant portion of the fault current in a system, even though the duration of the fault is normally short. An accurate prediction of this current is important to properly design the protection of the system, and to rate the circuit breakers. Many of the induction motors in use exhibit 'deep bar' behaviour, that is the referred rotor resistance and inductance change with changing frequency of the rotor currents. This paper examines the effect that the deep bar property has on the evaluation of the three phase symmetrical short circuit current. The deep bar model is verified against measured results of a short circuit applied to a 75 kW induction motor.

INTRODUCTION

When designing a new industrial reticulation scheme, it is very important to assess the fault levels of the system. Induction motors play a significant role in the peak fault levels that can be obtained which becomes important in determining rupturing capacities of circuit breakers and the setting of protection levels. The accurate determination of the fault level is important in terms primarily of safety and secondly in terms of capital cost. Often detailed fault studies are performed to determine the worst case fault which can be expected at different localities of the plant. In certain instances, for example if there is local generation, dynamic stability of the plant can also be an important consideration.

In many cases the fault current contribution of induction motors is calculated from simplified formula. Mulukutla and Gulachenski [1] show that the peak phase current can be approximated by

$$I_{pSC} = \frac{2V_p}{(X_1 + X'_2)} \quad (1)$$

DEVELOPMENT OF THE SHORT CIRCUIT CURRENT

In order to explain what happens to the induction motor when a three phase short circuit is applied to the terminals of the motor, it is useful to use the two axis model of the induction motor. Two important laws must be remembered while performing the analysis. Firstly the flux in the motor cannot undergo a step change, and secondly the rotor speed also cannot undergo a step change.

Consider the coils along the d axis of the motor, observed in a reference frame which is stationary with respect to the stator. Let the stator coil be fed from a voltage source, v , before the fault. The stator and rotor coils are mutually linked by the d axis flux. The rotor coil is short circuited on itself. An additional voltage rise occurs in the d axis rotor coil due to the fact that it is rotating with respect to the flux caused by the q axis coils; this voltage is termed a speed emf or the back emf. When the motor is operating at rated speed, the back emf e , is of a similar magnitude to that of the impressed stator voltage, and they also have the same frequency. The equivalent circuit of the d axis is shown in Figure 1. A similar circuit exists for the q axis.

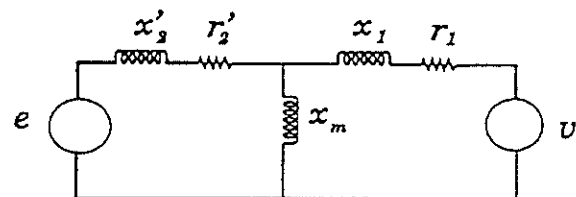


Figure 1. Equivalent circuit of the d axis

If the rotor quantities are referred to the stator, the voltage drops due to the resistances are neglected and the magnetising current is assumed to be small, an approximate equivalent circuit can be given by Figure 2. For the following analysis let us assume that the motor is running light, therefore the back

emf is almost equal to the applied voltage and hence the volt drop across the leakage inductances is small. The assumptions of this paragraph simply allow for an intuitive feel for what is happening when the motor is short circuited.

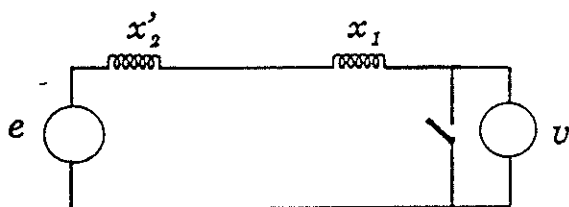


Figure 2 Approximate equivalent circuit.

Consider the case when the short circuit occurs when the d axis voltage is at a maximum value. At that instant, the maximum value of the back emf e , is impressed across the leakage inductances, $(x_1 + x_2')$. (Remember e is proportional to the q axis flux and the speed of the rotor, both of which do not change at the instant of the short circuit, therefore e exists at the same value after the short circuit as before it). The current flowing in both the stator coil and the rotor coil is given by the integral, with respect to time, of the voltage across the stator and rotor leakage inductances divided by the sum of these leakage inductances. Figure 3 (a) shows a typical trace of the current and the voltage. It is important to note that

- The current is evenly spaced about the zero axis.
- The magnitude of the current similar to that under direct on line starting, since the conditions are very similar. (When starting e is zero and v is impressed.)

Consider the case when the short circuit occurs when the d axis voltage is zero. The current flowing in the stator and the rotor coil is once again given by the integral of the voltage with respect to time divided by the sum of the leakage inductances. Figure 3 (b) shows a typical trace of the inductor current. It is important to note that

- The current has a DC offset equal to half the peak to peak current.
- The peak current is double that of the previous case, and is the same as given by equation (1).

For the case when the short circuit occurs for a different value of impressed voltage, a smaller DC offset will occur. Since the q axis is at quadrature to the d axis, at least one axis will have a DC offset in the current and in the general case both axes will have a DC offset.

This reasoning is also applicable to the actual phase currents of the motor, that is there will

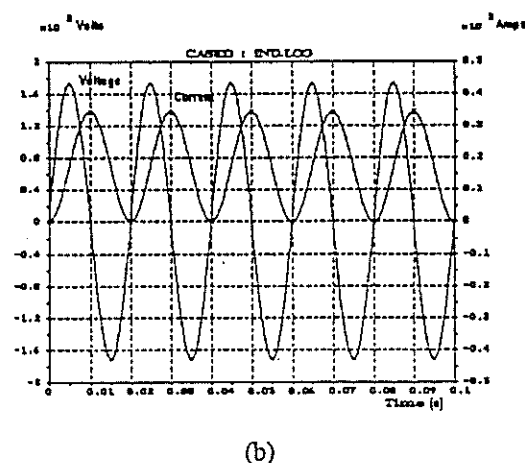
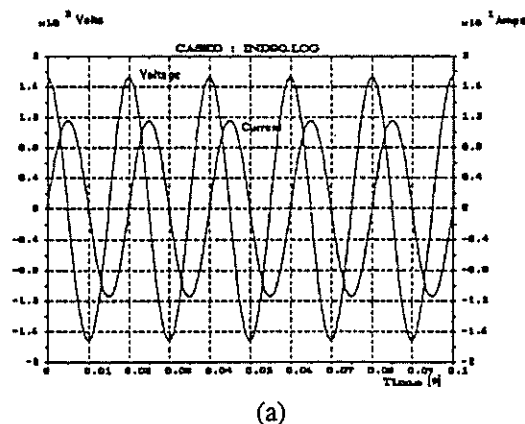


Figure 3 Voltage applied across an inductor and the resulting current
(a) Voltage starts at maximum value.
(b) Voltage starts at zero.

also be a DC offset in at least two of the three phases.

The effect of the resistances in the circuit is to decrease the both the AC component and the DC offset of the fault current. Further, as the energy stored in the main magnetic field is dissipated in the resistances, e will also decay, and hence the fault currents will decay with time. As the rotor slows down, the frequency of the generated emf e will also decrease.

The above analysis has been performed assuming that the currents are observed in a reference frame attached to the stator. However, it is important to see how the currents in the rotor are effected by the short circuit, when observed by in a reference frame which is attached to the rotor.

When moving from a reference frame fixed to the stator to a reference frame fixed to the rotor, the frequency of the currents are modulated by the rotor speed. This implies that a DC current observed in the stator reference frame is observed in the rotor reference frame as an AC current whose frequency is equal to that of the rotor speed multiplied by the pole pairs. Similarly an AC current, with a frequency dependant on the rotor speed, observed in the stator reference frame is observed in the rotor reference frame as a DC current, given that the AC current and the reference frame are rotating in the same direction.

From the above analysis it is clear that the rotor currents in an induction motor will contain two components of current just after the application of a three phase fault on the stator. The first component will be a decaying DC component, and the second will be a decaying AC component, whose frequency is proportional to the rotor speed. It is known that the effective impedance of the rotor changes with frequency, hence each of these two components will see a different impedance, which will in turn affects the peak stator currents. It is therefore important to examine the effect frequency has on the rotor impedance.

THE ROTOR INDUCTANCE

The leakage flux in a rotor bar caused by the current flowing in the rotor bar is shown in Figure 4. It is clear that all the leakage flux links with the portion of the bar furthest away from the air gap, and that very little of the leakage flux links with the portion of the bar closest to the air gap. The impedance of this flux is proportional to the frequency. If the frequency is zero, that is the current is DC, then the leakage flux produces no effect on the currents flowing in the bar. As the frequency increases, so does the impedance caused by the leakage flux.

Due to more flux linking with the portion of the bar furthest away from the air gap, the impedance to the current is larger here than closer to the air gap. The current therefore tends to migrate towards the air gap, where it is easier to flow. This current migration effectively changes the apparent shape of the bar, hence changing both the impedance and resistance of the rotor bars. This effect is often termed the deep bar effect. An analytical

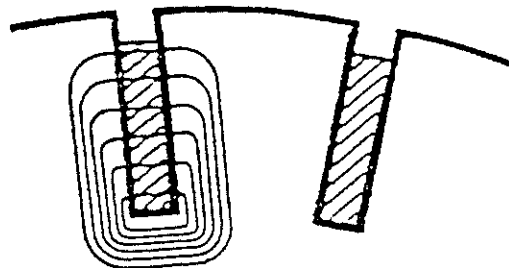


Figure 4: Distribution of the leakage flux in a rotor bar.

formula representing this phenomena has been presented by Alger [2].

From the previous section it was clear that the magnitude of the fault current was dependant on the value of the rotor leakage inductance. However it was also shown that the current flowing in the rotor bars consisted of a DC component and an AC component whose frequency was close to that of the supply frequency. The problem therefore arises in the choice of the rotor impedance, either for the DC current or the AC current. The formula developed by Alger does not allow for two currents of two different frequencies to flow simultaneously in the rotor bar, so another approach needs to be adopted.

IMPROVED MODEL OF THE ROTOR BAR

The problem of modelling the rotor bar in the presence of more than one frequency of rotor current has been examined by Levy et al[3]. The basic premise of that work is to represent the rotor bar as a very carefully chosen ladder network, as shown in Figure 5. The values of the inductances and the resistances are derived from the physical shape of the bar, and therefore require detailed information about the design of the motor. However excellent correlation of the starting currents was shown, and therefore this model was used to try and more accurately predict the short circuit currents.

COMPARISON OF SIMULATION OF FAULT CURRENTS

In order to show that the resulting short circuit current is dependant on the change in the rotor impedance, two simulations were performed of a three phase short circuit current being applied to a 600 kW induction motor. The first

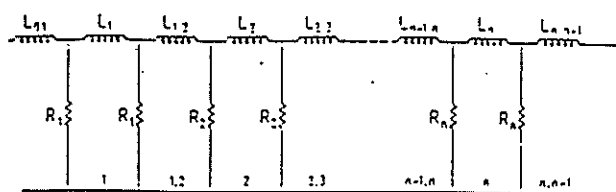


Figure 5: Ladder network representing a rotor bar.

simulation uses a model in which the rotor impedance is kept constant. The results are shown in Figure 6; the peak current is shown to be - 6 200 A (4.1 pu). The second simulation uses the model of Levy et al[3], which does account for the variation of the parameters in the rotor. The results are shown in Figure 7; the peak current is -10 050A (6.6 pu), an increase of about 60%! This example shows that the simple model does not adequately model the rotor bar and is inadequate for the evaluation of the short circuit current.

From Equation (1), this calculated peak current is 9010A for the model using the running parameters, and 9770A when using the starting parameters. It is clear that the latter provides an answer which is within 5% of the actual current.

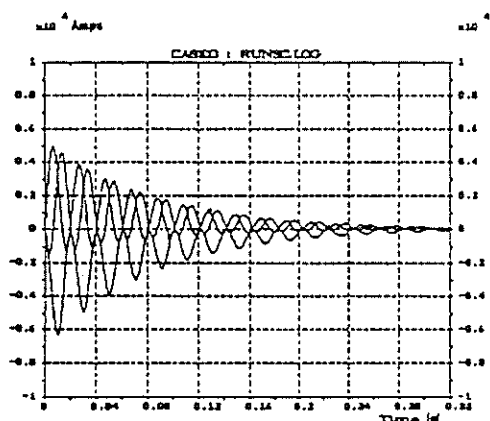


Figure 6: Simulation of the short circuit currents using conventional model.

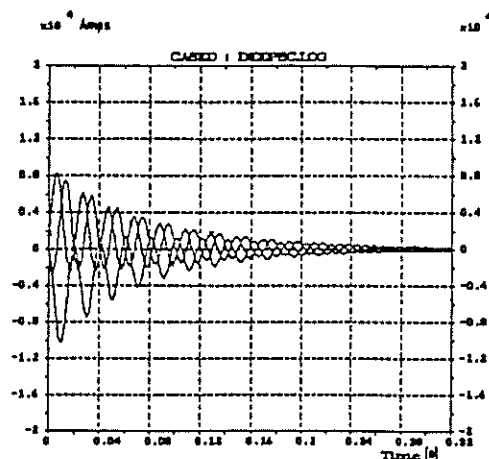


Figure 7: Simulation of the short circuit currents using improved model.

COMPARISON BETWEEN MEASURED AND SIMULATED RESULTS.

In order to verify the accuracy of the improved model, measurements were taken of a three phase short circuit applied to a 75 kW, 380V induction motor and compared to that of a simulation using the improved model. The induction motor was fed from a 400 kVA synchronous generator via a circuit breaker. The three phase short was created using knife switches connected between the lines feeding the induction motor. The motor currents were measured using Rogofski coils which are air cored and therefore do not saturate. When the fault was applied, the circuit breaker tripped, so as to disconnect the generator from the fault. This method therefore is a direct short circuit measurement, and not an indirect short circuit measurement.

The exact time at which the short is applied relative to the incoming voltage was not measured, so the results show the vector addition of the three currents, which is independent of the time of the fault. Figure 8 shows the predicted current, while Figure 9 shows the measured short circuit fault currents.

The following salient points can be seen when comparing the actual to the predicted fault currents:

- There is a very good correlation between the peak currents: 3400 A predicted, 3200 A measured.
- 10 ms after the fault, the measured currents underwent a shift in phase. One possible explanation is that the circuit breaker

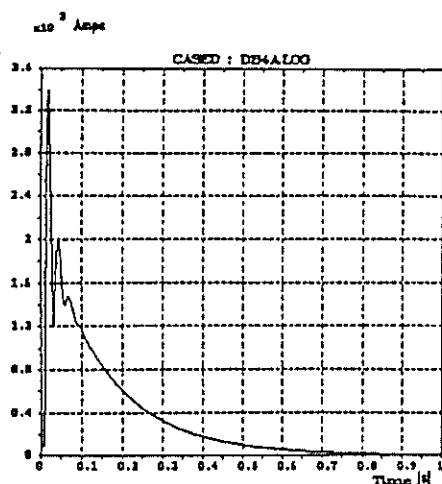


Figure 8: Simulation of the Short Circuit Current of a 75 KW Motor using the Improved Model.

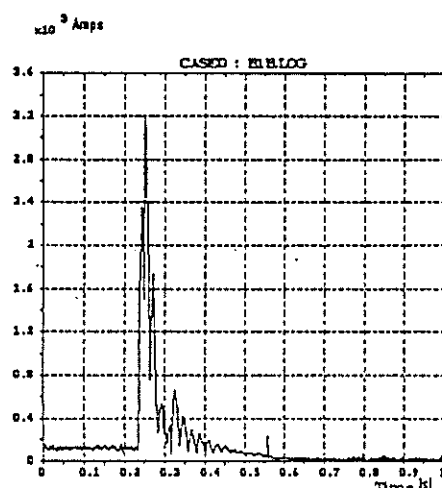


Figure 9: Measurement of the Short Circuit Current of a 75 kW Motor.

tripped at this point, causing this change. However this phenomena is still being investigated.

- The shape of the predicted and measured results show good correlation.

From these results, it is clear that the improved model does accurately predict the short circuit currents.

CONCLUSION

This paper has shown the development of the short circuit currents, and in particular the cause of the AC and DC components of current in both the stator and the rotor. The impedance of the rotor bar was shown to be different between the two components of the current because of the deep bar effect. It was therefore suggested that an improved model of the induction motor be used when predicting the short circuit current, especially if simulation was being used.

This paper has also shown that in order to calculate the short circuit current of an induction motor it is preferable to use the motor parameters applicable to starting conditions, because of the change in the frequency of the rotor currents to near supply frequency.

Measurements were taken on a 75 kW motor, which show that the improved model accurately predicts the short circuit currents.

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COMPUTATION OF INDUCED VOLTAGE AND SYNCHRONOUS REACTANCES IN PERMANENT MAGNET SYNCHRONOUS MACHINE

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ABSTRACT

The paper deals with a performance calculation of a particular design of permanent magnet synchronous machine. The machine under consideration is a synchronous machine with a solid steel rotor, suitable for high speed operation, excited by buried Samarium-Cobalt permanent magnets. The aim of this paper is to examine two existing methods of performance calculation, based on finite element (FE) magnetic field solutions, in terms of accuracy offered for this specific machine. The two quantities chosen for this comparison are stator induced voltage and synchronous reactances. A FE model of the machine is built and the magnetic field distribution calculated. Measured results from a prototype machine are compared with predicted results. Conclusions are drawn on the accuracy and applicability of each of the two methods used.

INTRODUCTION

Permanent magnet (PM) synchronous motor using high energy magnets are particularly well suited for high performance, variable speed drives. The specific application considered in this paper is a high speed PM synchronous motor having self starting capability.

A number of different possible rotor geometries can be considered for PM machines. For added strength the high speed rotor is sometimes surrounded by a cylinder or sleeve which may be magnetic or nonmagnetic. This paper describes a rotor which combines the requirements of high speed operation, i.e. robustness and smooth outer surface, with high air gap flux density and low leakage flux. The rotor of this machine (Fig. 1) consists of solid

steel instead of laminations and contains rectangular blocks of rare earth - cobalt magnets. To reduce the leakage flux round holes are drilled in the solid rotor body, in the spaces between the magnets, leaving thin bridges between the poles. This ensures that the leakage flux is significantly reduced (interpolar bridges saturate) while the surface of the rotor remains smooth. To accommodate the magnets four rectangular holes are machined using the spark erosion technique.

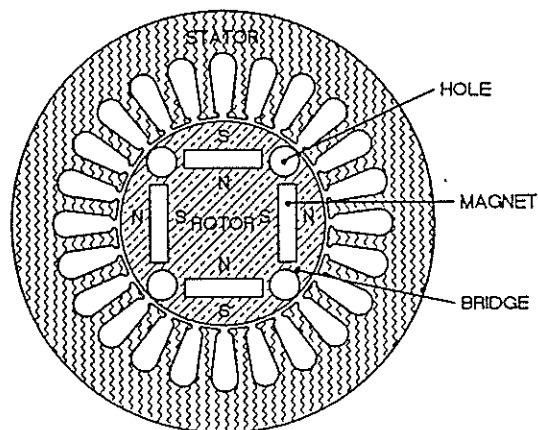


Fig. 1. Cross-sectional view of PM high speed machine.

OPEN-CIRCUIT VOLTAGE

Two different concepts (each one with two further variations) are used to calculate the open-circuit voltage of the machine. The first, called the flux density concept, utilises the distribution of the radial component of the magnetic flux density in the air gap. The second concept, called the flux linkage concept, is based on a change of the stator winding flux linkage with the rotor position.

Flux density concept

As a first step the distribution of the radial component of the magnetic flux density as a function of the circumferential distance along the air gap should be calculated using FE technique. A harmonic analysis of this waveform yields the magnitudes of the space harmonics present in the wave. The rms value E_v of each voltage time harmonic (or emf) generated in one stator phase by each flux density space harmonic is given by [1] :

$$E_v = 4.44 \nu f_1 k_{w\nu} N \frac{2}{\pi} B_v \frac{\tau}{\nu} l_s \quad (1)$$

where:

- ν = harmonic order
- f_1 = frequency of the fundamental
- $k_{w\nu}$ = winding factor for ν th harmonic
- N = number of turns per phase
- B_v = max value of the ν th harmonic of the radial component of the magnetic flux density
- τ = pole pitch for the fundamental wave along the centre of the air gap
- l_s = active length of stator stack

Finally, the resultant rms value of the stator emf per one phase is calculated as:

$$E = \sqrt{E_1^2 + E_3^2 + E_5^2 + \dots + E_k^2} \quad (2)$$

Since the procedure described above involves harmonic analysis and therefore is relatively complex, a simpler approach may be considered which assumes that the maximum value of the radial component of the air gap flux density $B_{\max}$ is equal to the magnitude of the first harmonic B_1 . This assumption may appear to be risky because it is clear that some harmonics of the air gap flux density wave, especially those related to the stator slots, can not be neglected. Indeed, this approach has proved to generate unacceptably inaccurate results.

Flux linkage concept

This method of calculating the open-circuit voltage is based on the following well known relationship between the induced voltage or emf e and the flux linkage ψ

$$e = \frac{d\psi}{dt} \quad (3)$$

The first step in the procedure is to obtain the stator phase flux linkages for two rotor positions. For this purpose the rotor is rotated by one slot pitch and the FE analysis is performed for each rotor position. The flux linkages are calculated for each stator slot separately and then combined to give the total flux linkage for one stator phase. The stator phase flux linkage ψ , obtained in this way, varies as a function of the rotor position θ . Assuming constant rotor speed Ω eq. (3) can be rewritten as:

$$e = \frac{d\psi}{d\theta} \frac{d\theta}{dt} = \Omega \frac{d\psi}{d\theta} \quad (4)$$

The instantaneous emf e is calculated using eqn. (4) and the rms value E of this emf is calculated as :

$$E = \frac{e_{\max}}{\sqrt{2}} = \frac{e}{\sqrt{2} \sin \theta} \quad (5)$$

where θ is the angle by which the rotor is rotated.

As an alternative quicker solution, the rotor movement may be simulated by moving the stator winding by the same angle (one slot pitch) but in the opposite direction. This requires only one instead of two FE analyses.

Results

The results of all four methods to calculate the open-circuit voltage are compared in the Table 1 with the actual open-circuit voltage measured on the prototype machine. The comparison reveals that both the flux linkage method and the method based on harmonic analysis of the air gap flux density give very good results. As expected, the flux density method which neglects the presence of the slot harmonics in the air gap flux density curve gives unacceptably poor accuracy and should not be used for machines having high slot opening-to-air gap ratio.

Taking into account both accuracy and calculation time, the flux linkage method which simulates the rotor movement can be recommended.

Table 1. Comparison of the measured and calculated open-circuit rms voltage at $n = 1500$ rpm.

calculated [V]				measured [V]
flux density method		flux linkage method		
all harm.		rotor turned	winding turned	—
32.0	36.2	31.2	31.3	29.9

SYNCHRONOUS REACTANCES

Similarly as for the open-circuit voltage two approaches based on different concepts are considered. The results are then compared and verified and the usefulness of both approaches discussed.

Stored energy method

The method of finding the synchronous reactance relies on the relationship between the inductance of a winding L and the energy stored in the magnetic field linked with the winding W_m :

$$L = \frac{2 W_m}{I^2} \quad (6)$$

where i is the current flowing through the winding.

The expression for the reactance X per phase of a three phase symmetrical winding has the following form:

$$X = 2\pi f \frac{2 W_m}{3 I^2} \quad (7)$$

where f is the rated frequency and I is the rms value of the winding current.

The stored magnetic energy is calculated by a FE package as:

$$W_m = \int_V \left(\int_0^B H \, dB \right) dV \quad (8)$$

Calculation of synchronous reactances X_d and X_q requires two different FE models. Stator MMF should peak in the rotor d axis for X_d calculation and in the rotor q axis for calculation of X_q . The calculated values of X_d and X_q as a function of stator current appear in Fig. 2.

Flux linkage method

The stator winding inductance L can be also defined as:

$$L = \frac{\psi}{I} \quad (9)$$

where ψ is flux linkage of the winding and I is the current flowing through it. This same definition was the basis of the method of synchronous reactances calculation described above.

The first step of the procedure is to perform a FE analysis for the two distinct cases: direct and quadrature axes of the synchronous machine. The FE models are arranged exactly in the same way as for the stored energy method described above.

The next step is to calculate flux linkages for all three phase belts. This is done as described previously in the section on the open-circuit voltage calculation.

Finally, the synchronous reactances of the machine are calculated using the following formulae:

$$X_d = 2\pi f \frac{\psi_d}{I_d} \quad (10)$$

$$X_q = 2\pi f \frac{\psi_q}{I_q} \quad (11)$$

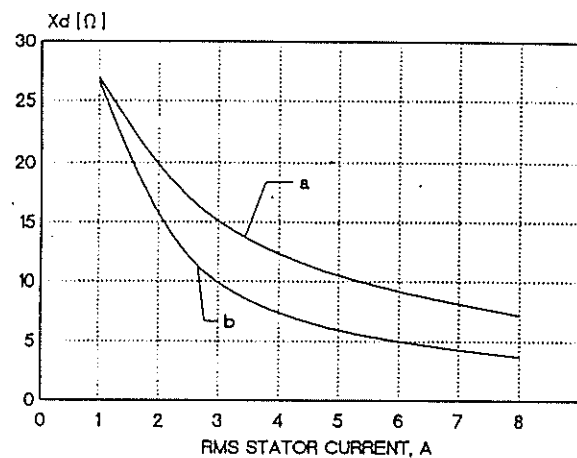
The calculated values of X_d and X_q plotted as functions of the stator current are presented in Fig. 2.

Comparison of results

The reactances calculated by the flux linkage method are, for saturated conditions, higher than those calculated by the energy method. This phenomenon,

characteristic for saturated magnetic circuits with small air gaps, is described and explained in detail in [2]. Briefly stated, this difference results from the fact that the two methods used are based on the different definitions of inductance, namely as stated by eqs. (6) and (9). These different definitions coincide for the linear case, as can be seen in Fig. 2 for non-saturated conditions (low currents), but lead to quite different values when the magnetic circuit saturates. To summarise up, the usefulness of both methods, especially for the machine with a short air gap, seems to be limited to low-saturation cases.

(i)



(ii)

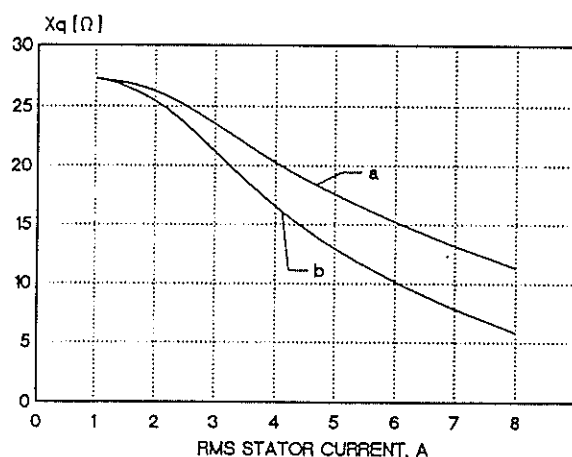


Fig. 2. Synchronous reactances as functions of the stator current : (i) direct axis, (ii) quadrature axis; a - flux linkage method, b - energy method

CONCLUSIONS

Selected variables of machine performance, namely the induced voltage and the synchronous reactances, have been calculated for a PM synchronous machine with a solid rotor and a small air gap. All the methods used have been based on a FE magnetic field solution.

Two different methods have been used to calculate each of variables and, where possible, the best method has been identified.

The FE approach has proven to be very suitable for this class of PM synchronous machines, however some limitations associated with a small air gap have been discovered.

ACKNOWLEDGEMENTS

The authors gratefully acknowledge financial assistance from the Foundation for Research Development and from the University of Natal

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POWER LINE FAULTS DUE TO GROUND FIRES: ELECTRIC FIELD CONSIDERATIONS FOR PARTICLE-INITIATED BREAKDOWN OF LONG AIR GAPS

by
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ABSTRACT

The likely factors involved in the electrical breakdown of long air gaps that have been invaded by carbonaceous products of combustion are briefly considered. Tentative conclusions are reached on some design aspects of power lines that cross sugar-cane fields.

1. INTRODUCTION

The electrical failure of the air between the conductors of 275 kV and 400 kV power lines in Natal is of major concern to Eskom and local industry. Such faults are due to sugar-cane burning and veld fires; of significance, the lower voltage lines are much less affected [1].

A feature of these power lines is that they have only two conductors per phase bundle (Figure 1) and so the electric field at the conductor's surface is fairly high. The questions have, therefore, been asked: (a) would an increase in the number of subconductors have a marked beneficial effect? (b) what increase in spacing between the conductors is necessary to avoid breakdown? The consideration of these questions is started herein, both theoretically and experimentally.

2. KEY FACTS

The air gaps of the affected lines are large; typical values are 8 to 9m between the phase conductors and about 5m from the phase conductor to the shield wire. Fire-initiated breakdowns mostly occur at mid-span; corresponding ground clearances are typically 8 to 9m. Of note: flashovers to ground are rare.

For clean air, the voltage required to cause such breakdowns is many times the normal operating value. Consequently, if normal service breakdown were to occur

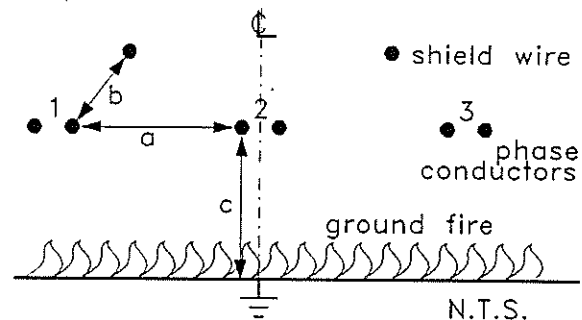


Figure 1: Typical conductor-fire situation.

solely due to thermally-reduced air density, then a very high temperature would be required; e.g. 1000°C for approximately a five-fold lowering of the electric strength. Although these fires are fierce, the air temperature at conductor level is only 120°C [2] to ~ 160°C [3] whilst the maximum flame temperature is 800–900°C [2]. Also, thermal ionization of air molecules is low – i.e. < 0.01% – for temperatures below 1000°C [4] and so it has a negligible effect on the breakdown strength.

Therefore, it is reasoned that particles are the main cause of the problem. To date, preliminary investigations have involved both small particles [2] and large ones [5].

3. GAS DISCHARGE ASPECTS

A conducting, or high permittivity, particle can be characterized by a stress enhancement factor, β , at its surface and a function, $f(z)$, that specifies the reduction in this field with distance, z , from its surface in the direction of the field. For an unperturbed field value E_0 , the corresponding field E_z is given as:

$$E_z = \beta f(z) E_0 \quad (1)$$

For a spherical particle of radius a

$$\beta f(z) = 1 + 2(a/a+z)^3 \quad (2)$$

Although the prolate spheroid can be dealt with analytically, the more irregular shapes are much more troublesome. Further, the influence of corona around sharply-pointed particles is a complicating factor. Therefore, a pragmatic approach is to assume:

$$\beta f(z) = 1 + (\beta-1) \exp -mz/l \quad (3)$$

where m is a constant and l is the characteristic length of the particle, which is half the actual length of a mid-gap particle and the full length of one that abuts the conductor. Previous investigations have shown that $m \approx 2$ and β has an effective value of 15 to 20 [5].

An electron, in being accelerated sufficiently over a distance d in the direction of the electric field, produces an avalanche of size n ; given as:

$$n = \exp \int_0^d \bar{\alpha} dz \quad (4)$$

where $\bar{\alpha}$ (mm^{-1}) is the effective ionization coefficient. For air at a temperature of 140°C , this coefficient is given as a function of the electric field, E (kV/mm), by the approximate relationship [4]:

$$\bar{\alpha} = 15500 \exp -21/E \quad (5)$$

From equations 1 through 5 and with $E = E_z$, the unperturbed field E_0 necessary to satisfy the streamer criterion of $n = 10^8$ electrons [4] is 1.5 kV/mm for spherical particles of 1mm radius and 0.13 kV/mm for thin particles of 300mm length (assumed $\beta = 18$).

4. OVERHEAD LINE FIELD

The three air gaps of particular interest (Figure 1) are: (a) phase to phase (b) phase to shield wire and (c) phase to ground. The determination of the electric field is facilitated by first considering an isolated phase and then generalizing to include the proximity effect of the other two phases and the shield wires.

4.1 Isolated Phase

From Gauss's theorem and the theory of electrical images, the potential ϕ at position x, y — in the plane of x axis along the line joining the conductors and y axis passing midway between them — is obtained from the general equation:

$$\phi = q(2\pi\epsilon)^{-1} \ln K^{-1} = V \ln 2rs(h^2K)^{-1} \quad (6)$$

where

$$K^2 = \frac{[(x-s)^2 + y^2][(x+s)^2 + y^2]}{[(x-s)^2 + (y+h)^2][(x+s)^2 + (y+h)^2]} \quad (7)$$

in which r is the conductor radius, $2s$ is the spacing between the subconductors, $h/2$ is the conductor's height above the ground, q is the charge per unit length of each conductor, V is the conductor's voltage, ϵ is the permittivity and $h \gg s \gg r$.

The single-conductor case can be dealt with by putting $s = 0$ whilst a fairly simple model exists [6] to cope with more than two conductors per phase.

4.2 Three Phases

For phases called 1, 2 and 3 and spaced horizontally at a distance S , the potential ϕ_{21} at phase 2 due to phase 1 is obtained from equations 6 and 7 by substituting $x = S, y = 0$ and noting that $S \gg s$:

$$\phi_{21} = q_1(2\pi\epsilon)^{-1} \ln(1+h^2/S^2) \quad (8)$$

where q_1 is the charge on phase 1.

A similar treatment applies for ϕ_{23} . Therefore, as $V = \sum \phi$, the values of q_1, q_2 and q_3 can be easily determined. Note: for practical cases, the contribution to the potential at the phase conductor due to the charges on the shield wires can be ignored (i.e. $<3\%$).

4.3 Shield Wire

The charges on the two shield wires — each of radius r_s , at spacing $2L$ apart and at height $H/2$ above ground — are readily determined by using equations 6, 7 and 8 by making $\sum \phi = 0$ and, where appropriate, putting $s = 0, h = H$ and $r = r_s$. This charge essentially determines the field close to its respective shield wire.

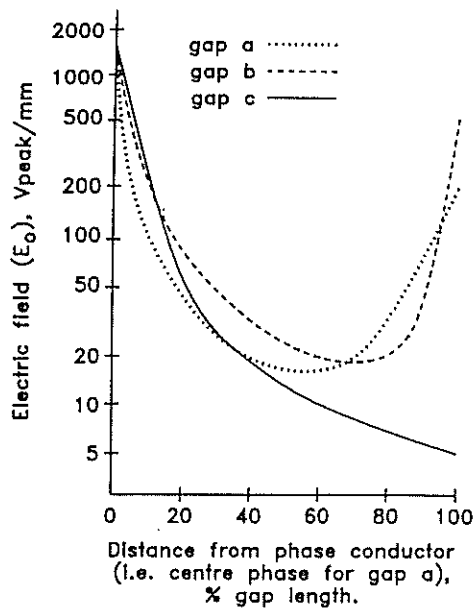


Figure 2: Unperturbed electric fields of 275 kV lines for air gaps specified in Figure 1.

4.4 Typical Values

From the computed potentials, the corresponding electric field, E_o , across the three specified air gaps a, b and c (Fig. 1) are presented in Figure 2 for 275 kV lines beneath which sugar-cane burning occurs. This field is specified as the peak value so as to be relevant to the gas discharge aspects. Such values have been computed using the data provided in Table 1 and at the instance in time giving the maximum value of the minimum stress in the gap; i.e. peak voltage on centre phase for gaps a and c and on a side phase for gap b.

5. EXPERIMENTAL

To complement the small-gap (i.e. 0.5m) study involving large particles [5], a few similar investigations have been made for a 5m gap and simulated overhead line conductors – horizontally supported 1.5m above ground. Twenty-seven particles, each a 0.3m long drawing pencil, were

Table 1: Values of parameters for calculations using equations 6, 7 and 8; at various system voltages.

System Voltage kV rms	Parameter						
	r mm	r_s mm	s m	S m	L m	h m	H m
132	9.0	4.4	–	6.0	4.5	15.2	25.2
275	14.3	6.6	0.2	7.9	7.1	15.7	25.2
400	17.8	6.6	0.2	9.3	8.3	17.5	27.4

Table 2: Voltages at which the first discharge is seen and the breakdown occurs, for various conductor and particle arrangements.

Conductor* arrangement at high-voltage and	Voltage, kV rms			
	First Discharge		Breakdown	
	HV-end particle**		HV-end particle	
	abutted	spaced	abutted	spaced
Single, 32mm dia	37	62	110	120
Single, 52mm dia	37	72	110	120
Twin, horizontal	38	64	118	115
Twin, vertical	46	79	120	120
Quad.	50	90	110	110

* For twin and quad arrangements, the subconductors were 32 mm diameter tubes spaced 0.4m apart.

** End particle either abutted or was spaced 20mm from the HV conductor.

equally spaced (i.e. 20mm between adjacent particles) in line along the direction of the maximum electric field. For each of the five arrangements – comprising single, twin or quad high-voltage conductor – the end-particle first abutted this conductor and then was spaced 20mm away. The earthed-conductor comprised a single 30mm diameter tube.

By slowly increasing the voltage, the first discharges to and between the particles started at the high-voltage conductor – because of the influence of the general earth. Upon the further increase in voltage, such discharging extended to more inter-particle gaps; when it involved about half the array, total breakdown occurred. The corresponding values of voltage are given in Table 2.

In a related experiment without particles, the breakdown voltage of a 300mm gap between the conductor bundle and an earthed plane was 103 kV rms and 114 kV rms for respectively the twin-conductor arrangement (conductors equally spaced from the plane) and the quad-conductor one (mounted as per the previous case).

6. DISCUSSION

6.1 Conductor Bundle

For N subconductors in a bundle symmetrically arranged on a pitch circle of radius R , the maximum electric field E_c at a subconductor's surface is derived from Cahen and Pelissier's model [6] as:

$$E_c \approx (2\pi\epsilon)^{-1} [q/r + (N-1)q/R] \quad (9)$$

where $Nq = Q$, the total charge on the phase conductor.

As N increases, Q increases; but as this rise is only logarithmic, E_c decreases. In changing from $N = 1$ (i.e. single conductor) to $N = 2$ (i.e. twin conductor) to $N = 4$ (i.e. quad conductor) but with other parameters remaining constant, the approximate ratio of E_c values is obtained from equation 9 as 1:0.54:0.31.

This substantial reduction in surface stress with increased number of subconductors explains the findings from the large-particle experiments with regard to the voltage at which the first discharge occurs (Table 2). In spite of this fact, however, the breakdown voltage was unaffected. Turning to detail, the influence of the end-particle's position on discharge inception is consistent with the corresponding results of the small-scale experiments and so has already been explained [5]. Corresponding deductions concerning a vast number of small particles are more difficult to make. A possible way round this difficulty may be to regard the stress level as being increased throughout the gap by a constant factor having a maximum value of between 3 (spherical particles, equation 2) to about 18 (pointed particles, equation 3). This concept can be regarded as being equivalent to a clean-air gap of much reduced dimension, for which the experiment reported herein showed that increasing the number of subconductors from 2 to 4 produced only a small increase (i.e. 10%) in the breakdown voltage.

6.2 Conductor Spacings

From the calculated field values provided in Sections 3 and 4, it can be deduced that gap breakdown must start at the conductors — as was the case in the laboratory experiment reported herein. In some respects, the main puzzling feature is the fact that breakdown does not occur to ground — even though the air density within the fire is only $\sim 25\%$ of the STP value. Possibly the particle density in this region is much lower than that at conductor level and/or the minimum value of the field, $E_o(\min)$, is important in such a cascade breakdown process. For the phase to phase gap, Figure 2 and related calculations show

that $E_o(\min) = 21, 17$ and 7.4 V/mm for 400, 275 and 132 kV lines respectively. To reduce $E_o(\min)$ to 7 V/mm for the higher voltage lines, the phase to phase spacings would have to be 23m and 16m for 400 kV and 275 kV respectively; the shield wire would have to be correspondingly raised by 7.5m for the 400 kV line and 5m for the 275 kV line. Among other things — eg. obtrusiveness, wayleave — the effect of such changes on the lightning performance needs to be assessed.

For highly divergent fields in "clean" air gaps, the mean breakdown field reduces as the voltage increases [4]. This aspect, therefore, needs to be investigated for "dirty" gaps before a definitive viewpoint can be reached concerning such spacings.

7. CONCLUSIONS

Fire-initiated breakdowns across the air gaps of power lines are unlikely to be reduced appreciably by increasing the number of subconductors. The dominant parameter in this complex breakdown process may be the minimum value of the unperturbed electric field. To effect a substantial reduction in such events, the interconductor spacing may need to be 2 and 2.4 times the current values for 275 and 400 kV lines respectively; further investigations are necessary to determine precise data for redesign purposes.

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POLLUTION SEVERITY MONITOR FOR RELEVANCE TO INSULATOR FLASHOVER: SOME DESIGN FEATURES OF ONE BASED ON THE PELTIER EFFECT

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ABSTRACT

The concept is described of using a thermoelectric cooler to enhance the condensation of atmospheric moisture on an insulator surface to provide a pollution severity monitor based on electrical conductance. The findings from a trial arrangement based on the theoretical aspects explained in this paper are sufficiently encouraging to warrant a development programme.

1. INTRODUCTION

Power companies need to know the pollution severity of a location before their outdoor high-voltage insulators can be properly designed. Further, the pollution deposited on an installed insulator requires to be measured - or, at least, well estimated - in order to implement a cleaning programme.

Although numerous methods of making such measurements exist [1], none is ideal. In many respects, there is much to commend a technique based on the electrical conductance of the polluted surface. The main problems are that dry pollution is essentially insulating and the occurrence of natural wetting - e.g. high humidification, fog, dew, drizzle - is neither frequent nor controlled.

One solution is to surround the polluted insulator with a highly-moist atmosphere, e.g. a steam chamber [2]. Another solution is to cool the surface so as to enhance moisture deposition; the only such monitor reported to date employs a

conventional refrigerator feeding cooling coils built into an epoxy resin insulator [3]. The design of a more sophisticated method, based on the Peltier effect, is examined herein. Its foreseen advantages are that it would be compact, relatively cheap and provide information fairly quickly. Further, it would avoid the pollutants being leached from the insulator surface - which is particularly important if pollution severity needs to be continuously monitored. As electrical conductance can be readily measured when the local relative humidity at the insulator surface reaches 80% [4], this value can be the design criterion for the control circuitry of such a monitor.

2. PELTIER JUNCTION

In 1834, Jean C. A. Peltier discovered that the passage of an electric current through a junction consisting of two dissimilar metals can either cool or heat this junction depending on the direction of the current. This phenomenon is termed the *Peltier effect*. Figure 1 illustrates a

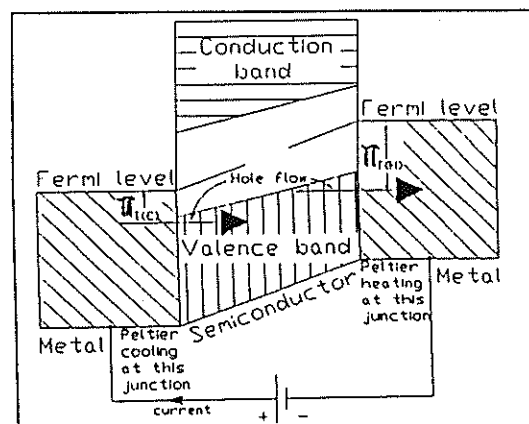


Figure 1. A junction of two dissimilar materials experiencing a current.

semiconductor connected between two metal contacts. The contacts are ohmic in nature and a dc power source is connected across the sample.

Peltier cooling (or heating) may be interpreted as being due to the change in the average kinetic energy of the current carriers when they cross a contact [5]. Note: the Fermi level in the metal is continuous with the Fermi level in the semiconductor when equilibrium exists across the interface; also, the position of the Fermi level in the metal is in the centre of the electron energy distribution in the partly filled band. When hole-current flows from left to right across the junction, it carries on average only the Fermi energy of that metal. For it to cross the left hand junction, however, it must have sufficient energy to overcome the Fermi barrier to get into the conduction band. Thus, only those holes in the metal at the high end of the distribution of kinetic energy can cross this contact and so heat is supplied by the surroundings to restore the average kinetic energy of the population of holes that remain behind.

The Peltier coefficient, π , gives the magnitude of the heating or cooling that occurs at a junction over and above the ordinary Joule heating. As π is the energy liberated or absorbed at the contact per coulomb of charge crossing that contact, it has the dimensions of voltage.

A common misconception is that a *Peltier* device somehow absorbs heat and carries it away - perhaps with the current. This is not true; the device only transfers or "pumps" heat from one of its sides to the opposite side. At the hot side, the heat must be removed through using a sink.

Figure 2 illustrates a configuration where many junctions are connected thermally in parallel and electrically in series; such a

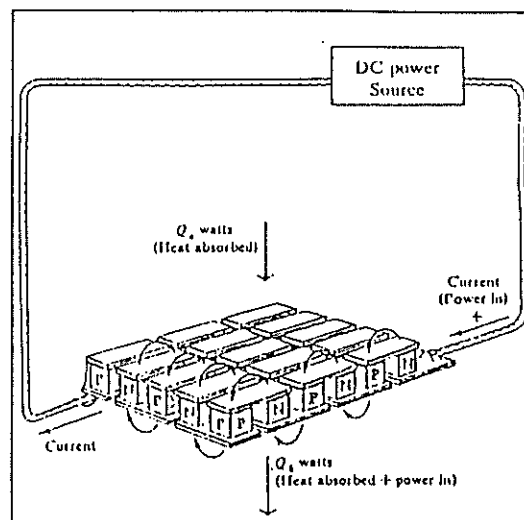


Figure 2. A thermoelectric cooler.

device is called a *Thermoelectric cooler*. The Peltier junction used for the pollution monitor could be of this type.

3. TEMPERATURE CONTROL

Relative Humidity (RH) is defined as the ratio of the actual vapour pressure, p , to the saturation vapour pressure, p_s , over a plane liquid surface at the same temperature; ie, in percentage terms:

$$RH = \frac{p}{p_s} \times 100\% \quad (1)$$

Saturation temperature (or dew point temperature) is the temperature at which a given mixture of water vapour and air is saturated; ie, for the pollution monitor application, it is the temperature of the air local to the insulator when minute specs of water appear on its surface.

A straightforward method of obtaining the relevant information is to use a simple wet-dry bulb thermometer arrangement. This item can be simulated electronically with temperature sensor IC's, of which one is covered with a wet wick. The respective information is best represented by means of a *Psychrometric chart* - as is illustrated in Figure 3.

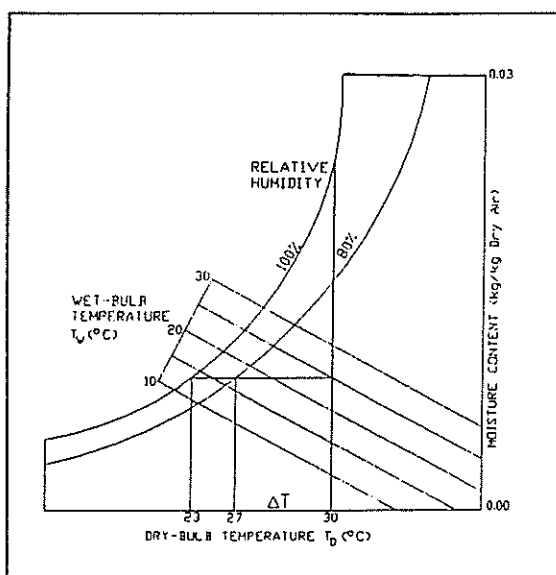


Figure 3. Simplified Psychrometric chart.

The Dry-bulb temperatures are read off the x-axis while the Wet-bulb temperatures are represented by diagonal lines (of negative gradient) which correspond to constant-enthalpy plots. The hyperbolic lines correspond to the relative humidity. Hence, if the wet-bulb and dry-bulb temperatures are known, the corresponding relative humidity is where the two lines intersect.

Therefore, if a horizontal line is drawn such that it passes through the point at which the wet and dry temperature lines intersect, it is then possible to determine the required temperature change, ΔT , to achieve the desired degree of saturation (relative humidity). Note: the vertical axis on the chart represents absolute humidity. Further, it should be appreciated that this absolute humidity remains unchanged as the dry-bulb temperature, T_D , is varied in order to obtain a specific relative humidity.

The Psychrometric chart defining equations can be manipulated such that the respective 80% relative humidity temperatures are software-generated and

programmed into an Eprom (and ultimately be micro-processor driven) in the form of a "look-up" Table.

3.1 Defining Equations

The basic equations required to ascertain the saturation temperature from the wet and dry bulb temperatures are the Wet-bulb equation and Antoine's equation [6].

The Wet-bulb temperature is the dynamic equilibrium temperature attained by a water surface when the rate of heat transfer to it by convection equals the rate of mass transfer of water away from that surface due to evaporation. For the present purposes, the wet-bulb equation is written in terms of humidity differences as:

$$Y_s - Y = \frac{h}{\lambda} \times (T_D - T_w) \quad (2)$$

where, Y = actual humidity; Y_s = saturated humidity; h is the heat-transfer coefficient = $1.6 \text{ kJ/kg} \cdot ^\circ\text{C}$; T_D and T_w are the dry bulb and wet bulb temperatures ($^\circ\text{C}$), respectively. Humidity (Y and Y_s) may be defined as mole of H_2O /mole of air $\approx$ partial pressure of H_2O /1 atm; λ is the latent heat of vaporisation = 2454 kJ/kg at 20°C .

The Antoine equation provides a relationship between temperatures and pressures for saturated conditions. The equation has the form:

$$\log P = A - \frac{B}{C + T} \quad (3)$$

where P = saturated vapour pressure (kPa); T = temperature @ P ($^\circ\text{C}$); and for air, $A = 7$, $B = 1650^\circ\text{C}$, and $C = 226^\circ\text{C}$.

3.2 Calculating the 80% Saturation Temperature (T_{80})

When the atmospheric pressure remains constant, the saturated vapour pressure is determined from Antoine's equation for a known value of T_w . Equation 3 is thus rearranged to provide the saturated vapour pressure $P_{w\ sat}$ as:

$$P_{w\ sat} = 10^{7 - \frac{1650}{226 + T_w}} \quad (4)$$

It is now convenient to work with $P_{w\ sat}$ in units of atmospheres ie, $P_{w\ sat}(\text{Atm})$. Then $P_{w\ sat}(\text{Atm})$ is substituted into equation 2 to give the partial pressure of water vapour $P_w(\text{Atm})$ in atmospheres as:

$$P_w(\text{Atm}) = \frac{1.6}{\lambda}(T_w - T_D) + P_{w\ sat}(\text{Atm}) \quad (5)$$

For the 80% saturation temperature to be obtained, $P_w(\text{Atm})$ must be divided by 0.8 and multiplied by 101 (ie the conversion factor for changing atmospheres into kPa) to give P_{80} as:

$$P_{80} = \frac{101}{0.8} \times P_w(\text{Atm}) \quad (6)$$

P_{80} is then substituted back into Antoine's equation so that T_{80} can be determined from:

$$T_{80} = \frac{1650}{7 - \log P_{80}} - 226 \quad (7)$$

The temperatures obtained from the *wet-dry bulb thermometer* and the Psychrometric chart agreed to within 3% of those from a Vaisala-HMT31 temperature/RH probe.

4. THE INSULATOR

Due to the awkward shape and the physical size of commercial insulators, it

seems likely that a representative insulator should be used for convenience and ease of cooling. A trial *insulator* has been spun out of aluminium and its surface coated with electrically insulating varnish. It has been experimentally confirmed that such a surface is hydrophillic.

5. DISCUSSION

A trial arrangement consisting of *insulator*, Peltier junction and wet-dry bulb circuitry has been tested. While the wet-dry bulb sensor produced reasonably accurate ($\pm 5\%$) results, it is evident that this item would not be practical for extended field use. Therefore, once the experimental setup has been refined in the laboratory, a more appropriate humidity sensing device will have to be developed before proceeding to a service evaluation.

6. CONCLUSION

Although further development of the Peltier-based pollution monitoring device is necessary, the prospects look promising. Correlation between it and other pollution measuring methods, however, still need to be investigated.

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EMC IN THE CONTEXT OF A LARGE POWER UTILITY

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ABSTRACT

The paper discusses how the scope of electromagnetic compatibility, as historically applied to radio and telecommunications systems, is extended to large electric power systems which act as both sources and receptors of electromagnetic interference. It is shown that lightning remains the predominant source of natural interference. Broad research needs and activities in the Eskom context are briefly described.

1. INTRODUCTION

The field of Electromagnetic Compatibility (EMC) is the technological discipline which is concerned with the electrical effects of what is known as Electromagnetic Interference (EMI).

Historically, EMC has grown out of the need to limit EMI from electronic systems, and very importantly, to ensure that such systems are electrically compatible with each other. The focus has thus been on radio and telecommunications systems. In recent years, however, the growing use of electronic equipment in electric power systems - comprising generation, transmission, distribution and reticulation plant - has meant that electric power utilities must consider EMC in the design and operation of their networks. If this is not properly done, the performance of a system - in terms of the availability and continuity of supply of electrical energy - can be seriously affected.

This paper reviews the scope and significance of EMC in electric power systems, with emphasis on the power system itself as a major source of EMC. Of interest to the Saupec Conference will be the review of the EMC topics which require some local research.

2. WHAT IS EMC?

Electromagnetic compatibility (EMC) is the ability of electric equipment and systems to function without degradation or malfunction in their intended operational electromagnetic environment. The equipment or system should not adversely affect the operation of, or be affected adversely by any other equipment or system (1).

EMC is achieved when the **immunity** or **susceptibility** of a given system to impinging **emissions** of EMI from an adjacent system or source are adequately reconciled or controlled. Emission and susceptibility levels thus have to be **co-ordinated** such that the interacting systems can operate in a technically compatible manner.

EMI is any unwanted signal, noise or field that produces damage to or an unwanted response in, a given piece of equipment or system. The unwanted disturbance may cover the frequency range from DC to light (1).

For any given EMI problem, there are three necessary elements as shown in figure 1.

The source of EMI may be **natural** or **man-made**. The transfer medium refers to one in which **conduction**, **coupling** or **radiation** of interference may take place. Susceptible receptors are found in the following categories of systems or equipment :-

- Industrial scientific and medical equipment
- Electric power systems
- Communications and radio systems
- Business machines and systems.



Figure 1 : Basic Elements Of Any EMI Problem

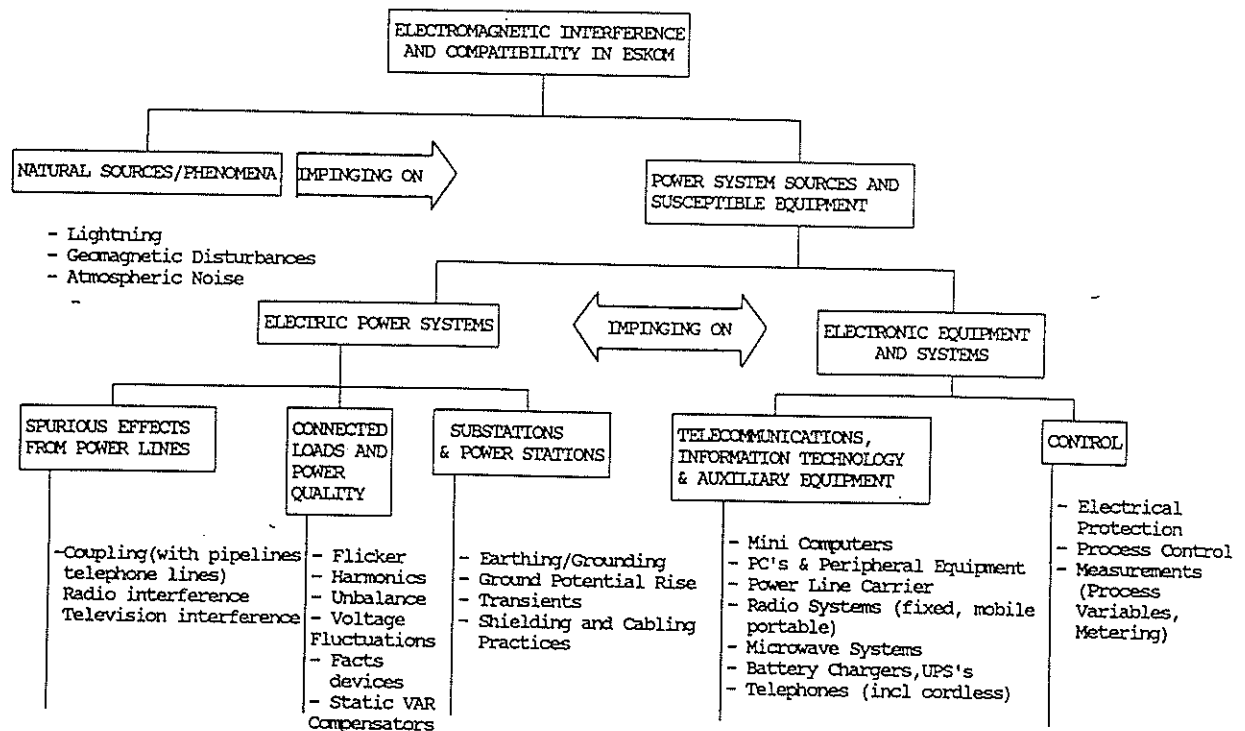


Figure 2: Interactions between sources and receptors of electromagnetic interference in large electric power systems.

3. SCOPE OF EMC IN A LARGE POWER SYSTEM

Figure 2 shows in a summarised form the various sources of EMI and their interaction with systems used by a large power utility such as Eskom.

Natural Sources

The main natural source of EMI which has a great impact on power systems is lightning. For many decades the prevention of equipment damage caused by lightning has been a primary design and operational consideration in most electric power systems. Lightning can, however, also act as a source of interference to sensitive electronic equipment in substations and power stations (3).

Geomagnetic disturbances, whilst a major problem to electric networks in parts of Canada and Sweden, are not, as far as is known, disruptive to networks in this part of the Southern Hemisphere.

Man-made sources

The power system itself contains various EMI sources and susceptible equipment. The main sources, comprising emissions from power lines, substations and connected loads, may interfere with the electronic

equipment used to control the network. This equipment may also cause interference to adjacent electronic systems, irrespective of interference coming from the network itself. Malfunctions in such equipment can in turn cause malfunctions in substation or power station control systems; this can cause loss of generation and interruptions of supply.

Emissions from power lines (radio interference, for instance) have an influence on local radio reception. Such emissions have to be limited to levels compatible with the local environment. For this reason, careful attention has to be given to the design of high voltage lines.

Voltage distortion caused by connected loads is also a source of EMI, in a broad sense. The levels of harmonic current emission from large non-linear loads must meet strict limits so that the resulting voltage distortion at the point of connection does not adversely disturb other equipment in the network. This also applies to loads which cause voltage unbalance, flicker and voltage fluctuations. Harmonics, unbalance, flicker and voltage fluctuations (which may also originate in the supply network itself) contribute to what is known as "voltage quality" or "quality of supply". This in itself is a very specialised subject, and is being intensively addressed by several large utilities. It is thus a major subset of EMC in any large power system.

Faults in power systems are also a major source of EMI, giving rise to voltage transients within a substation, and voltage surges or depressions in the wider network. The resulting transients impinging on substation control equipment constitute a major disturbing source of EMI. Susceptible equipment (protection relays, energy meters, computers) require rigorous immunity testing - mainly in accordance with the IEC - 801 series (4). Switching within a substation (breakers, isolators) is also a source of radiated and conducted interference.

Electronic equipment within the power station, substation and control centres must not only withstand the disturbances imposed on them by lightning or the power system itself, but also EMI from adjacent electronic systems. The CISPR recommendations (which deal mainly with limits of emission from various types of equipment) are often applied to ensure that EMC is achieved (7).

A particular concern in power stations and substations is EMI from portable radio transceivers and, lately, the use of cordless phones near sensitive equipment. Radio transceivers are known to have caused equipment malfunctions, whilst, so far, little is known about the compatibility between cordless phones and nearby electronic systems.

EMC Standards

The major means by which a utility can promote and to a large degree ensure EMC in its plant is to enforce the application of relevant EMC standards and recommendations. As noted, CISPR covers limits of radio frequency emission from electronic equipment; immunity to transient disturbances and steady state radio frequency emissions are covered by the IEC-801, 555 and 1 000 series of specifications (4, 5, 6).

4. AREAS OF FUTURE STUDY AND RESEARCH

Recent reviews of EMC problems done by Eskom have shown that research is required in a number of areas. Such research can be divided into three categories :-

- (i) Surveys and assessment of the EMI environment in Eskom installations.
- (ii) Research into specific aspects which are strongly influenced by local factors, but which are not being researched elsewhere.
- (iii) Assessment of commercial software for modelling disturbances (lightning and fault induced transients, and harmonic distortion) in power station and substation plant.

All three categories must be underpinned by the maintenance of adequate local engineering, computational and analytical skills in EMC.

It is not the intention to provide precise details of proposed and evolving research activities here; what follows is the basis for further analysis and evaluation.

Surveys

1. The growing use of integrated protection and measurement systems in the newer substations - and also those being refurbished - has resulted in the need to measure and assess the levels of radiated, conducted and coupled interference generated by isolator switching in large substations. This work is about to be started by Eskom.
2. Distributed control systems (such as long ash conveyors) at several large power stations have periodically experienced lightning induced damage and maloperations. How wide and generic the problems are is not yet known; this requires a detailed survey to be done. Some work in Eskom is about to be started on this aspect.
3. In terms of quality of supply, a survey of harmonic, unbalance and flicker distortion levels, together with the occurrence of voltage dips, is about to be started by Eskom. The main aim of the survey is the quantification of distortion levels, and from this, the provision of a factual basis for better management of deleterious effects in Eskom's transmission and distribution networks.

Specific Studies

1. As cordless telephones (46 and 900 MHz band equipment) are becoming available for use in power stations and substations, it is necessary to assess the immunity of protection and control systems to emissions from such equipment. Little research has been done elsewhere, as far as is known. Eskom has already started some studies on this topic.
2. Eskom makes extensive use of static var compensators (SVC's) in various 400kV networks. Such devices generate high levels of electromagnetic noise which enter on operational power line carrier circuits. The judicious choice of carrier frequencies has so far avoided major interference problems, but there is a need to improve the modelling of the propagation of this noise within future substations for which SVC's are being planned.

3. The emergence of FACTS devices means that additional sources of EMI will possibly be present in future transmission systems. The impact of this form of EMI on substation control systems thus requires initial study.

Software Assessments

Several software packages for modelling the dynamic response of substation earthing systems during faults and lightning disturbances are now commercially available. These require careful evaluation. Furthermore, surveys need to be done of the availability and suitability of other software for modelling wave propagation from lightning ground flashes and subsequent coupling with electric power systems.

Maintenance And Development Of Expertise

As should be clear from this review, EMC in a power system context is a complex topic - as such it should be supported by good EMC engineering and research skills. As regards research, high level skills in measurement, analysis and numerical computation (computational electromagnetics) are indispensable, and will contribute to solving the foregoing problems.

5. CONCLUDING REMARKS

This paper has been of a review nature, and has examined a broad range of topics in EMC, albeit superficially in the case of specific research needs. The paper does, however, show the following :-

- (i) EMC in a power system environment is broader in definition and scope than that historically associated with radio and telecommunications systems.
- (ii) Disturbances caused by lightning have always been, and will remain, a major source of EMI in power systems.
- (iii) Loads connected to a power system can be major sources of EMI, notably as sources of current and consequent voltage distortion.
- (iv) Disturbances in the power system itself (faults and switching) are sources of EMI. Electronic measurement and control systems need to be immune to such disturbances, ranging from voltage dips and transient overvoltages to steady state and transient radio frequency disturbances.
- (v) Research in the form of (i) EMI surveys (ii) specific studies, for example EMI modelling and EMI from cordless phones, and (iii) software evaluation is required.

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WERKVERRIGTINGSIMULASIE VAN 'n HIBRIEDE VOERTUIG

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Opsomming - In hierdie referaat word eerstens 'n kort beskrywing gegee van hibriedevoertuie, hulle moontlike konfigurasies en watter rol hulle moontlik in die toekoms mag speel. 'n Oorsig word daarna gegee van die komponente waaruit die gesimuleerde voertuig saamgestel is en faktore wat in ag geneem is by die simulase daarvan. Laastens word simulasiresultate gegee met die klem op die verband tussen die vermoë van die drywing lewerende komponente en die energiekoste tydens 'n gegewe rit.

1. INLEIDING

Sedert die sestigerjare het battery aangedrewe voertuie hernude aandag begin kry. 'n Nadeel van hierdie voertuie is egter die beperking wat relatiewe lae energie tot massa verhoudings van hedendaagse batterye daarop plaas. Hierdie probleem het hibriede voertuie as 'n alternatief op die voorgrond gebring. (Vir die doel van hierdie referaat word met die woord *hibriede voertuie* verwys na 'n voertuig waarvan die aandryfstelsel bestaan uit 'n binnebrandenjien en batterye.)

Die motivering agter die ontwikkeling van hibriede voertuie met die oog op die bekamping van lug sowel as geraas besoedeling het die afgelope tyd drasties toegeneem. Veral in die lig van besoedelingswetgewing wat in ontwikkelde lande 'n algemene verskynsel geword het.

Sover dit hibriede voertuie aangaan is daar basies een van twee ontwerpfilosofieë wat gevolg kan word. Onderskeid word hier getref deur te bepaal watter een van die energie bronne as primêre en watter een as die sekondêre bron aangewend word. In hibriede voertuie word ook meestal van die elektriese motor as 'n generator gebruik gemaak tydens die remproses ten einde die batterye te herlaai en sodoende energie te herwin.

Daar bestaan twee alternatiewe stelsel konfigurasies wat gekarakteriseer word deur die wyse waarop drywing vanaf die energiebronne na die voertuig se as oorgedra word. Dit kan volkome elektries (seriekonfigurasie) of elektro-meganies gekombineerd (parallelkonfigurasie) wees. In die geval waar dit volkome elektries is word meganiese energie vanaf die binnebrandenjien eers na elektriese energie omgeskakel voordat dit aangewend word om die voertuig aan te dryf.

Die doel van die ondersoek was om 'n praktiese hibriede aandryfstelsel te modelleer, en dan die verband

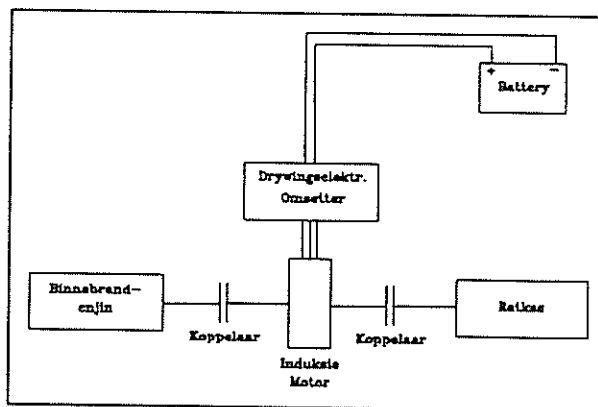
vas te stel tussen die grootte van die elektriese motor en die energiekoste, wanneer die voertuig op stedelike ritsiklusse gebruik word.

2. DIE SIMULASIEMODEL

2.1 Keuse van hibriede samestelling vir simulase doeleindes

Daar is besluit om vir die doeleindes van die simulase en gepaardgaande ondersoek op 'n spesifieke samestelling van 'n hibriede voertuig te besluit. Die kombinasie waarop besluit is, is gebaseer op 'n model van Volkswagen wat op die Geneefse motorskou van 1990 vertoon is.

Die voertuig is gesamentlik deur Bosch en Volkswagen ontwikkel en die hibriede aandryfstelsel is in 'n standaard Volkswagen Golf gemonteer. Die hibriede aandryfstelsel vorm 'n enkel-as sisteem en bestaan uit 'n 1.6 liter, 42 kW Volkswagen katalitiese diesel enjin wat op 'n as aan 'n 6 kW, 72 V induksie motor / generator en 'n vyf spoed handratkas gekoppel is. Dit vorm dus 'n parallelkonfigurasie en word in Figuur 1 getoon.



Figuur 1: Model van VW se hibriede voertuig

2.2 Benadering tot die Simulasie.

Voordat enige simulase kan plaasvind is dit noodsaaklik dat die stelsel wat gesimuleer moet word, sowel as die omstandighede waaronder die stelsel funksioneer volledig gedefinieer is i.t.v. stelsel- en omgewingsparameters.

In die geval van 'n hibriede voertuig sal die stelselparameters in die eerste plek die dinamiese eienskappe van die voertuig beskryf (bv. windweerstand, rolweerstand, voertuigmassa ens.) en in die tweede plek die vermoë van die voertuig

omdrywing te kan lower.

Tydens die simulase is die voertuig beskou as 'n stelsel wat 'n sekere drywingsuitset moet lewer. Die oorsprong van die drywing is die twee energiebronne nl. die batterye en die brandstofenk. Drywing word dus langs 'n roete van verskillende komponente na die wiele van die voertuig oorgedra (sien Figuur 1) met gepaardgaande verliese in elke komponent.

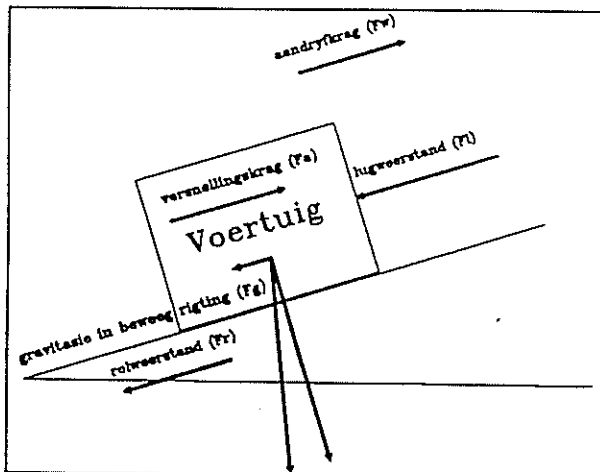
'n Wiskundige formulering van elk van die betrokke verliese was dus nodig ten einde energieverbruik te kan bepaal. 'n Uiteensetting van hoe die drywing wat benodig word bereken is, tesame met verliese wat in berekening gebring is, word in die volgende paragrawe gegee. Wiskundige detail van bg. is weens 'n gebrek aan ruimte beperk. Inligting hieroor is in die verwysings beskikbaar.

2.3 Berekening van drywing wat by die wiele verlang word

Al die betrokke kragte word in Figuur 2 getoon en die krag wat die voertuig se wiele op die pad uitoefen is as volg:

$$F_w - F_g - F_l - F_r = F_a \quad (1)$$

$$F_w = F_a + F_r + F_l + F_g \quad N$$



Figuur 2: Kragte op voertuig

Elk van bogenoemde kragte word deur fisiese wette beskryf. Die drywing wat by die wiele benodig word is as volg:

$$P_w = F_w \cdot V \quad W \quad (2)$$

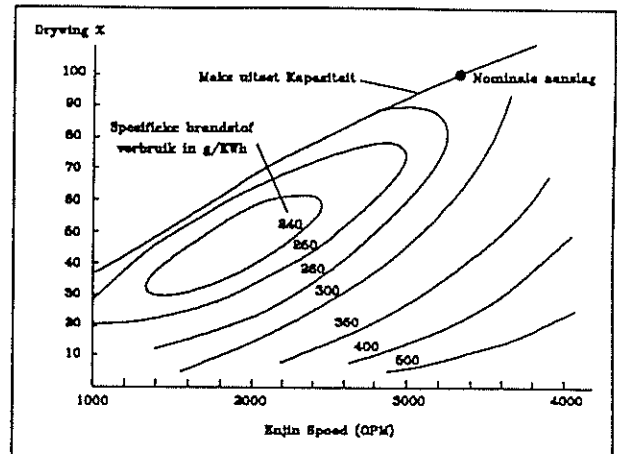
waar P_w die drywing by die wiele is en V die snelheid van die voertuig.

2.5 Berekening van verliese in ratkas

Ratkasverliese kan in twee komponente verdeel word: eerstens spoedafhanklike verliese weens die olie, ratte en laers, tweedens lasafhanklike verliese, wat eweredig is aan die drywing wat oorgedra word. Laasgenoemde kom voor weens wrywing tussen die ratte.

2.5 Berekening van verliese in die binnebrandenjnin

Ten einde die brandstofverbruik van die binnebrandenjnin te bepaal is daar van 'n genormaliseerde enjinkaart wat die spesifieke brandstofverbruik by verskillende werkpunte gee, gebruik gemaak. So 'n genormaliseerde enjinkaart vir 'n dieselenjin word in Figuur 3 getoon.



Figuur 3: Tipiese genormaliseerde brandstofverbruik kaart van 'n dieselenjin

Aangesien die energie inhoud van diesel bekend is, is dit moontlik om verliese in die binnebrand enjin te bepaal.

2.6 Bepaling van verliese in die induksiemotor

(a) Koperverliese

Die koperverliese is suiwer 'n weerstandsverlies en is dus afhanklik van die grootte van die strome in die rotor- en statorwikkels van die induksiemotor.

(b) Spoed-afhanklike verlieskomponente

Daar is hoofsaaklik twee spoed-afhanklike verlieskomponente nl. die verliese in die laers van die motor en die windverliese. Die wrywingsverliese in die laers verloop lineêr met spoed terwyl die windverliese kwadrates met spoed toeneem. [4]

(c) Swerfverliese

Swerfverliese (stray losses) is moeilik definieerbaar, maar vir die doel van hierdie simulase was dit voldoende om dit as 'n verlies te beskou wat kwadrates met stroom sowel as frekwensie in die motor verloop [5].

(d) Kernverliese

Die aanname is gemaak dat die kern binne sy lineêre gebied op die histerese lus bedryf word. Onder hierdie omstandighede sal die kernverliese lineêr met spanning sowel as frekwensie toeneem.

2.7 Drywingsverliese in GS na WS omsetter

Die tipe omsetter wat gemodeleer is staan bekend as 'n volbrug pulswydtemodulasie (PWM) gelykstroom na wisselstroom omsetter. Hierdie GS-WS omsetter laat ook drywingsvloei in die teenoorgestelde rigting toe. Die verliese in so 'n omsetter kan hoofsaaklik in twee komponente verdeel word nl. skakelverliese en aanverliese.

Skakelverliese is afhanklik van die styg- en valtye, asook die grootte van strome deur en spannings oor die skakelkomponente. Aan-verliese is ook afhanklik van die strome en spannings deur en in die transistors sowel as die persentasie van die tyd wat die stroom vloei. [6]

2.8 Effek van filter tussen battery en GS na WS omsetter

Aangesien 'n battery se drywingsdigtheid nadelig beïnvloed word wanneer 'n diskontinue stroom daaruit getrek word [1], is besluit om 'n LC-filter op die inset van die GS-WS omsetter te gebruik. Die verliese in so 'n filter kan in die geval as weglaatbaar beskou word. [6]

2.9 Bepaling van die battery se verliese

Verliese in die battery is te wyte aan die interne weerstand van die battery. Hierdie weerstand neem toe namate die battery ontlai.

'n Ander belangrike faktor ontrent batterye wat nodig was om te modelleer, is die feit dat die energievermoë van 'n battery afhanklik is van die tempo waarteen dit ontlai word [2]. Verder is daar ook 'n afname in klemspanning van batterye namate dit ontlai [3].

3. TOEPASSING VAN DIE MODEL

3.1 Formulerings en berekening

Die verwerkliking van die simulatie was heelwat meer kompleks gewees as net die toepassing van 'n reeks wiskundige formules.

Die rede hiervoor is die interafhanklikheid tussen die verskillende komponente. Die feit dat die uitset van een komponent dien as inset vir 'n volgende, sowel as 'n gebrek aan analitiese oplosmetodes vir sekere komponente se stel vergelykings (bv. die induksiemotor) het die nodigheid van verskeie iteratiewe berekeninge vereis.

3.2 Ondersoek wat m.b.v die simulatie gedoen is

Die doel van die ondersoek was om die verband tussen die grootte van die induksiemotor wat gebruik word, en die totale koste van die energie wat verbruik is, m.a.w. die koste om die brandstof tenk weer te vul en om die batterye weer te herlaai, te bepaal. As sekondêre doel is die effektiwiteit van die induksiemasjien en die

dieselenjin onder hierdie spesifieke omstandighede ook ondersoek.

3.3 Energieverdeling-strategie

In die ondersoek is die benodigde drywing so ver as moontlik deur die induksiemotor gelewer. Ten einde egter binne die termiese vermoëns van die motor te bly is die maksimum uitset op enige tydstep beperk tot 80% van die maksimum drywing wat die induksiemotor kan lewer. Wanneer dit nodig word om hierdie perk te oorskrei is die dieselenjin ingespan.

Tydens negatiewe versnellingsperiodes word die induksiemasjien as generator benut om sodoende energie na die batterye terug te voer. Weereens word drywing in die induksiemasjien tot 80% van die maksimum vermoë beperk.

3.5 Voertuig eienskappe

Al die voertuigparameter is volledig gebaseer op die VW model wat vroeër vermeld is. Die enigste verskil is dat die induksiemasjien-grootte nie vasgestel is nie, maar gewissel word van 1 kW tot 10 kW. Die totale batterymassa is 215 kg.

3.6 Ritsiklus-spesifikasie

Die ritsiklus wat gebruik is, is 'n standaard siklus wat vir die toets van elektriese voertuie gebruik word nl. SAE J227a. Hierdie roete-profiel word i.t.v. snelheid vs. tyd in Figuur 6 getoon. Die padhelling en windsnelheid is as nul geneem.

Die aantal kere wat die roete herhaal is, is bepaal deur die aantal kere wat die voertuig die roete met 'n 6 kW motor kan herhaal (6 kW omdat dit die grootte motor is waarmee die VW model toegerus is) totdat die batterye 80 % ontlai is. Daar is gevind dat die roete 18 keer herhaal kan word, wat 'n afstand van 23,6 km impliseer.

4. RESULTATE

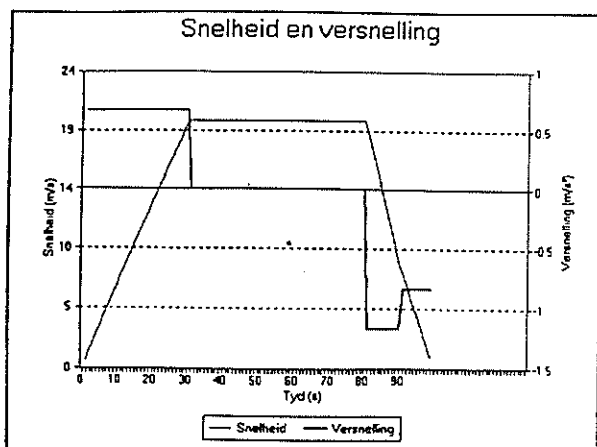
Omdat die waardes van alle voertuigparameters, asook die ritprofiel varieer kan word, is dit moontlik om 'n groot aantal scenarios te ondersoek.

Slegs enkele voorbeelde van die belangrikste resultate word egter in die figure hieronder getoon.

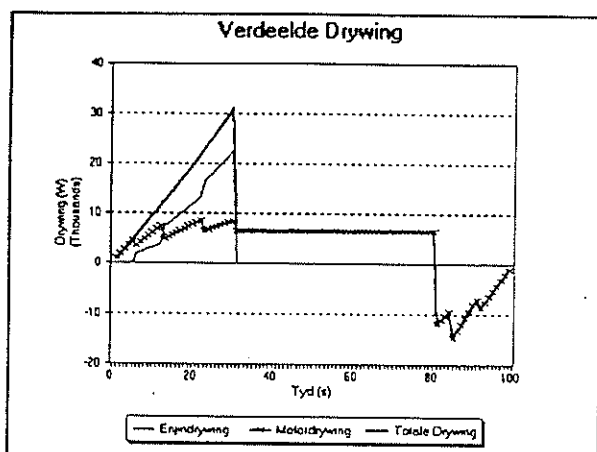
5. GEVOLGTREKKINGS

'n Realistiese model is geskep vir 'n hibriede diesel/elektriese voertuigaandryfstelsel, soortgelyk as wat op eksperimentele basis deur VW en Bosch gebou is.

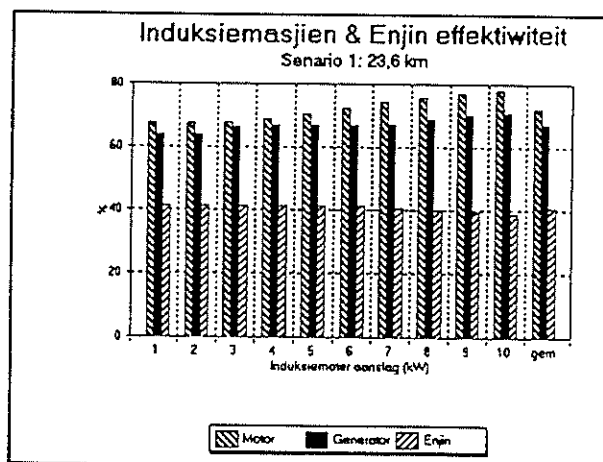
Die model is gebruik om die effek van komponentgroottes op die energiekoste van die voertuig te bestudeer, en insiggewende resultate is verkry



Figuur 4: Snelheid en versnelling profiel van SAJ227 skedule D roete

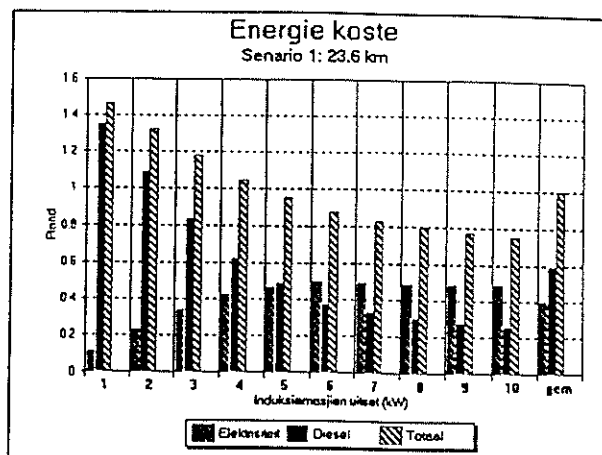


Figuur 5: Hier kan gesien word tipies hoe die gevraagde drywing tussen die induksiemotor en die enjin verdeel is.



Figuur 6: Effektiwiteit van die verskillende grootte induksie masjiene (as motor en generator) en die binnebrand enjin.

Die grootste nut van die model lê daarin dat ander parameters soos batterygrootte en -tipe, ritpatrone, padhellings ens. varieer kan word, ten einde die effek daarvan te bestudeer, en 'n aandryfstelsel vir 'n spesifieke doel te optimaliseer.



Figuur 7: Elektriese, diesel en totale energie koste

6. BRONNELYS

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DIENSKWALITEIT EN BEDRYFSKOSTE ANALISE OP 'N RETIKULASIESTELSEL

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1) Opsomming

'n MARKOV-model is ontwerp waarmee die bedryfskoste asook die kwaliteit van 'n stadsraad se diens aan huishoudelike verbruikers ondersoek kan word. Voorsiening is gemaak vir die toepassing van lasbeheer om energiekostes te verlaag en oorbelading van kables en transformators te voorkom. Deur lasbeheer toe te pas is dit moontlik om die opgradering van die verspreidingsnetwerk te vermy.

2) Inleiding

Die beskikbaarheid van elektriese energie aan huishoudelike verbruikers word deur verskeie faktore beïnvloed. Falings a.g.v. verswakking van die netwerk asook oorbelasting van kables en transformators kan voorkom. Verskeie beheerstrategieë kan ook toegepas word om die lasfaktor te verbeter, en oorbelasting te voorkom, deur die beskikbare elektriese energie aan die verbruiker in te perk. Energiekostes word sodoende afgebring en falings a.g.v. oorbelasting teëgewerk. Al dié faktore asook twee beheerstrategieë is d.m.v. 'n MARKOV model geïmplementeer. Die produk is 'n instrument waarmee dienskwaliteit asook bedryfskoste bepaal kan word vir verskeie lasbeheer opsies.

3) Waarom 'n MARKOV model?

'n Groot deel van 'n munisipaliteit se elektriese energie word versprei na huishoudelike verbruikers. Die verspreiding van hierdie energie geskied d.m.v. transmissielynne, transformators, kables, skakeltuig, en oorhoofse geleiers. Die kompleksiteit en grootte van die verspreidingsnetwerk neem gevolglik toe, terwyl daar nader beweeg word na die eindverbruiker. Aangesien die netwerk van geleiers, transformators, kables ens. blootgestel word aan die natuur en die omgewing, kan die stelsel nie onbepaald funksioneer sonder om iewers te faal nie. Dit gebeur bv. dikwels dat 'n kabel raakgekap en beseer word deur 'n persoon wat 'n sloot grawe. Kables en transformators het ook 'n beperkte drywingsvermoë wat nie veilig oorskry kan word nie. Indien die vermoë van 'n kabel of transformator oorskry word, sal die beveiliging in die stelsel die betrokke kabel of transformator klink. Dieselfde kabel wat gefaal het a.g.v. skade, "faal" nou omdat dit oorbelas word. Ons sien dus dat die beskikbaarheid van elektriese energie aan 'n huishoudelike verbruiker beïnvloed kan word deur twee tipes falings. Die eerste tipe noem ons vervolgens algemene falings en die tweede tipe oorbeladingsfalings. In beide falingsgevalle

word die toevoer na die verbruiker tydelik onderbreek terwyl die fout herstel word. Dit is egter nie altyd nodig om die toevoer te onderbreek tydens 'n fout nie. 'n Alternatiewe kabel kan bv. gebruik word terwyl die beskadigde kabel herstel word. 'n Naasliggende transformator kan ook bv. ingespan word om tydelik energie aan verbruikers te voorsien terwyl 'n beskadigde transformator herstel word. Aangesien die naasliggende netwerke die hoofverspreidingsnetwerk kan ondersteun, bied dit die voorsieningsowerheid 'n geleentheid om sy diens te verbeter.

Die verskeidenheid van tariewe waarmee die voorsieningsowerheid (stadsraad) deur ESKOM aangeslaan kan word, bied ook geleentheid i.t.v. kostebesparing d.m.v. lasbeheer. Die model is ontwerp om verskeie lasbeheer opsies vir verskillende tariewe toe te pas. Die bedryfskoste asook dienskwaliteit van elke opsie kan met mekaar vergelyk word om sodoende die mees doeltreffende strategie te implementeer.

Om die kwaliteit van 'n voorsieningsowerheid se diens aan sy kliënte, asook die koste van die diens te bepaal, was 'n model nodig om die beskikbaarheid van die verspreidingsstelsel te simuleer. 'n MARKOV-model wat lasbeheer, algemene falings, oorbeladingsfalings, alternatiewe toevoere, instandhouding reaksietyd en kostes suksesvol in ag neem, voldoen aan hierdie behoefte. Die produk is 'n instrument waarmee die voorsieningsowerheid sy diens aan verbruikers kan optimeer.

4) Hoe lyk die model?

Die verspreidingsnetwerk tot by Lynnwood-Mews in Lynnwood, Pretoria word in dié bespreking beskou. Die stadsraad van Pretoria versprei energie deur die stad d.m.v. 'n ring van 132 kV transmissielynne. By die hoofsubstasie in Lynnwood word hierdie 132 kV omgeskakel na 11 kV. 'n 11 kV kabel loop vervolgens vanaf die hoofsubstasie deur die gebied terug na die hoofsubstasie. Dié kabel word d.m.v. 'n skakelaar onderbreek om twee radiale te vorm, maar kan gebruik word om enige transformator van albei kante af van energie te voorsien. Gevolglik word die kabel gedefinieer as 'n ring. Op verskeie punte langs die ring word energie afgetap d.m.v. transformators wat 'n uitset van 415 V (240 V) aan die huise lewer. Oorhoofse geleiers verbind hierdie transformators met die huise en, vir die doel van hierdie bespreking, word die oorhoofse geleiers gedefinieer as 'n lokale netwerk. Om te verseker dat die model aanvaarbare en realistiese inligting sal weergee en die werklikheid goed sal simuleer, sonder om die probleem te

kompleks te maak; is besluit om slegs die verspreiding vanaf die 132 kV:11 kV transformator na die huise te beskou.

Die verspreidingsnetwerk kan dus deur die MARKOV-model gesimuleer word wat in figuur 1 aangetoon is. Die model bestaan uit die volgende moontlike toestande:

- [1] Normaal: Energieverspreiding normaal soos ontwerp is.
- [2] Kabel faal:
Kabel of skakeltuig het gefaal, energie word d.m.v. bystandskabel voorsien
- [6] Kabel bystand faal:
Die alternatiewe kabelroete het gefaal.
- [3] Transformator faal:
Die transformator of lokale laespanningsnetwerk het gefaal. Die bystandsnetwerk voorsien krag aan verbruikers.
- [7] Transformator bystand faal: Die bystandstelsel waarmee verbruikers met energie voorsien word i.g.v. 'n transformator fout het ook gefaal.
- [4] Ring beheer:
Die normale energiepatroon word beperk a.g.v. lasbeheer wat die las op die kabel monitor.
- [8] Ring beheer faal:
Die ring is oorbelaas en het die kabel uitgeklink.
- [5] Transformatorbeheer:
Die normale energiepatroon word beperk a.g.v. lasbeheer wat die las op die 11 kV:415 V transformator monitor.
- [9] Transformatorbeheer faal:
Die ring is oorbelaas en het die transformator uitgeklink.

5) Hoe werk die model?

'n Beheerstrategie waarmee die maksimum aanvraagtarief (ESKOM Tarief A) optimaal benut word, asook 'n strategie wat die nuwe tyd van gebruik tarief (ESKOM Tarief T2) optimaal benut, is d.m.v. die model ondersoek. Lasdata wat gemeet is by die 11 kV substasie, word in die model ingelees en aangepas om te voorspel wat die las op 'n 11 kV:415 V transformator in die stelsel is. Lasmettings word sodoende tot 'n minimum beperk. Die model gebruik nou die ingeleesde laskromme en die een wat ontstaan het, om die beheerstrategie by twee verskillende punte (Hoofsubstasie (132 kV:11 kV) en transformator (11 kV:415 V)) te simuleer. Die maksimum aanvraagstrategie beperk die energie gelewer tot 'n maksimum grenswaarde deur sekere laste af te skakel. 'n Tyd van gebruik strategie daarenteen skakel alle voorafbepaalde laste (bv. waterverwarmers) tydens die spitsye af om soveel geld moontlik te bespaar. Die MARKOV model maak gebruik van maklik bekombare inligting om die falintempo's asook hersteltempo's vir elke uur in die dag uit te werk voor die stogastiese matrikse saamgestel en die D.V.'s opgelos word. Die definisies wat gebruik is in die bepaling van bedryfskoste asook dienskwaliteit word gegee in

vergelykings (1),(2),(3),(4). Aangesien die huishoudelike verbruiker sommige laste meer belangrik ag as ander, is weegfaktore in die definisie van dienskwaliteit gebruik om die kostebesparing tydens lasbeheer af te speel teenoor die hoeveelheid energie wat geskuif moet word om die beparing teweeg te bring. Lasbeheer kan bv. gebruik word om energiekostes te bespaar sonder om die verbruiker te frustreer deur waterverwarmers aan en af te skakel. Die kostes wat sodoende bespaar kan word weeg in hierdie geval swaarder as die belangrikheid van die energie wat verskuif word tydens lasbeheer.

6) Resultate verkry m.b.v. die model:

Die las in 'n huishoudelike gebied sal oor 'n tydperk van bv. tien jaar groei terwyl die energie-behoeftes van die verbruikers toeneem. Die voor-sieningsowerheid kan die netwerk opgradeer of lasbeheer begin toepas, indien die las die vermoëns van die gevestigde verspreidingsstelsel begin oorskry. Die model is gebruik om dié twee opsies teen mekaar af te speel. Realistiese data wat verkry is met die hulp van die Stadsraad van Pretoria is gebruik om drie verskillende scenarios te simuleer. Eerstens is die huidige stelsel gesimuleer waarna die huidige las verdubbel is. Laastens is die huidige laspatroon gebruik in die simulaties terwyl die stelsel se kapasiteit halveer is. Lasbeheer is toegepas deur 1 kW waterverwarmers aan en af te skakel. Die resultate word in tabelle 1,2 & 3 aangetoon.

7) Bespreking van resultate:

Indien die stelsel se vermoë met die helfte verminder word of indien die las verdubbel, sal die kabel en transformator beide oorbelaas word. Deur lasbeheer in hierdie gevalle toe te pas, kan verseker word dat die stelsel nie klink nie. Vir die hele ring word 'n kostebesparing tussen 7.23% en 7.98% voorspel met die toepassing van maksimum aanvraagbestuur. By die transformator kan 'n besparing van minstens 5.62% voorspel word met die toepassing van dieselfde bestuurstrategie. 'n Groter besparing kan egter by die transformator verwag word aangesien die aanpassing tussen die ring- en transformatorlaskromme slegs 'n konserwatiewe voorspelling van die werklikheid lewer. Die T-tariefstrategie lewer 'n besparing van $\pm 8\%$ afhangend van die aantal laste wat tydens spitsyd afgeskakel kan word. Met die huidige tyd van gebruik tariewe bestaan daar nie op die oomblik meer moontlikhede i.t.v. energiekostebesparing as met die maksimum aanvraagtarief nie. Die toepassing van 'n tyd van gebruik-beheerstrategie sal in werklikheid die stadsraad se diens verswak aangesien te veel las afgeskakel moet word om dieselfde energiekostebesparing as met die maksimum aanvraagstrategie teweeg te bring. In al die gevalle word bedryfskoste verminder deur lasbeheer toe te pas. Die dienskwaliteit van die stelsel verbeter aangesien:

- i) Die stelsel meer beskikbaar is.
- ii) Kostes bespaar word d.m.v. lasbeheer.

Die diens is swakker met die toepassing van 'n tyd van gebruikstrategie aangesien die verbruiker se energiegebruik te veel ingeperk word.

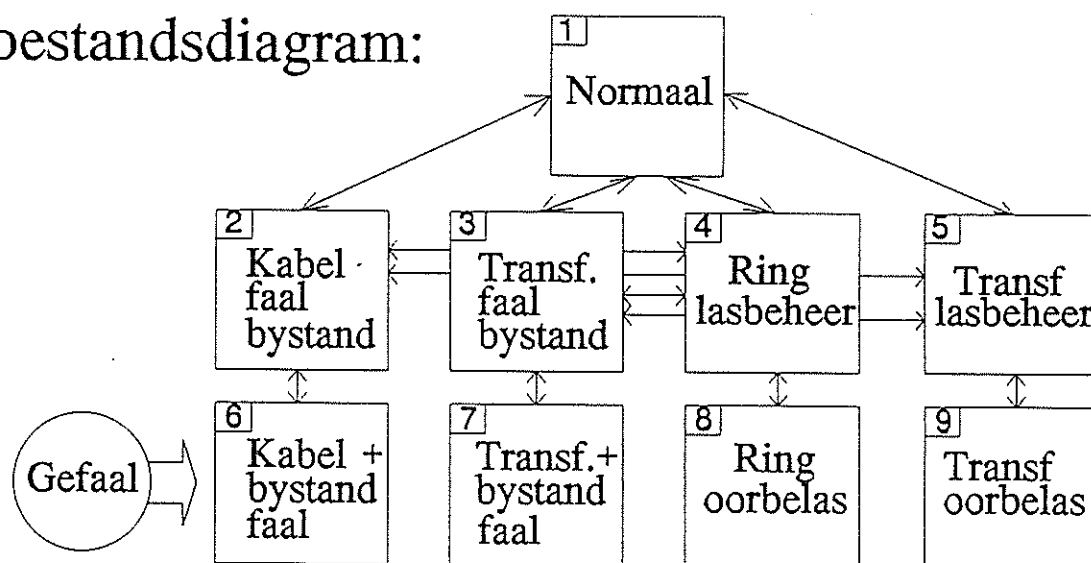
7) Bevindings:

Dit is duidelik dat die las se groei d.m.v. goeie lasbeheer ingeperk kan word. Selfs 'n las 2.5x groter as die huidige las, kan d.m.v. lasbeheer hanteer word. Maksimum aanvraagbeheer kan ook suksesvol toegepas word om energiekostes te bespaar. Die gevolge is 'n verbetering in dienskwaleit. Die tyd van gebruiktariewe bied egter nie op hierdie stadium dieselfde moontlikhede as die huidige maksimum aanvraagstariewe nie. Dit is duidelik dat lasbeheer die beste toegepas kan word indien die beheerder so globaal moontlik gestasioneer is. Sodoende word die stelsel se laskromme verbeter en word alle beheeraksies gekoördineer. Lasbeheer kan ook toegepas word om komponente in die stelsel (soos bv. die transformator) te beskerm. Opgraderings kan dus in die netwerk vermy word deur lasbeheer so na moontlik aan die probleem punt toe te pas.

Bronnelys:

- (1) Billinton, R., Allan, R.N.: "Reliability evaluation of engineering systems: Concepts and techniques." Longman Scientific & Technical
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Toestandsdiagram:



Figuur 1: MARKOV model

- Bedryfskoste = Koste van gewone bedryf
 + verlies in verkope a.g.v. falings
 + koste van herstel
 + energiekoste
 + koste om lasbeheerder te bedryf
 - besparing in energiekoste a.g.v. beheerstrategie. (1)
- Kostekomponent = (Bedryfskoste na implementering van strategie)
 /(Bedryfskoste van normale bedryf sonder strategie). (2)
- Energiekomponent = (Aangevraagde energie deur verbruiker tydens beheerperiode)
 /(Energie werklik gelewer tydens beheerperiode). (3)
- Dienskwaliteit = (Kostekomponent * Kosteweegfaktor + Energiekomponent * Energieweegfaktor) (4)
 *(1-Alle toestande waar absoluut geen energie beskikbaar is nie).

Tabel 1: Resultate van bedryfskoste analise: Ringbeheer

Bedryfskoste	M.A. Tarief		T-Tarief	
	Gewoon	Strategie	Gewoon	Strategie
Tans	R 1 264 317	R 1 172 215	R 1 412 596	R 1 308 837
½ Vermoë	R 1 264 317	R 1 172 215	R 1 412 596	R 1 302 973
2 x Las	R 2 635 254	R 2 424 972	R 2 946 641	R 2 454 297

Tabel 2: Resultate van bedryfskoste analise: Lokale beheer

Bedryfskoste	M.A. Tarief		T-Tarief	
	Gewoon	Strategie	Gewoon	Strategie
Tans	R 176 419	R 166 507	R 213 710	R 207 216
½ Vermoë	R 185 781	R 176 877	R 224 985	R 218 491
2 x Las	R 369 781	R 360 123	R 430 007	R 377 124

Tabel 3: Resultate van dienskwaliteit analise

Diens- kwaliteit	Gewoon	Ring beheer		Lokale beheer	
		M.A. Tarief	T.Tarief	M.A. Tarief	T-tarief
Tans	1	1.0240	0.9289	1.0066	0.9675
½ Vermoë	1	1.0240	0.9331	1.0028	0.9693
2 x Las	1	1.0193	0.9661	0.9994	0.9852

HIERARCHICAL STRUCTURES FOR INTEGRATED MINE ENERGY SYSTEMS PLANNING

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ABSTRACT

Considering the fact that 17% of all electrical energy consumed in South Africa is used by deep-level mines, it is of the utmost importance to develop a mine energy systems planning model to optimise the energy consumption of deep-level mines.

A basic conceptual hierarchical structured model for integrated mine electrical energy planning is being developed to accommodate the following factors:

- external influences on the system, such as electricity tariffs.
- internal influences on the system, such as system and process limitations and constraints and management requirements.
- change of energy carriers, such as changing from compressed air rock drilling to hydraulic rock drilling.

One of South Africa's biggest mining companies is supporting this development by making one of their shafts available for the experimental calibration of the model.

1 INTRODUCTION

Deep level mining in South Africa utilises about 10% of the electrical energy generated on the African continent [1]. A demand-side management program targeted at deep level mining offers significant opportunities for modification of the demand profile on the interconnected system in South Africa. The results of an initial investigation on a typical deep level mine have provided a better understanding of the demand-side management potential. Environmentally and socially responsible approaches to the cost-effective end-use of electricity in deep level mines are being exploited. These new approaches will stimulate the implementation of least-cost planning in the mining and electric utility industries in Africa.

In 1985, 27.4% of Eskom's energy sales went to the mining industry [2]. In 1989 it was 24% and the deep level gold mines consumed more than 17%. These mines are the largest single electricity consumer grouping in South Africa [3]. It is evident from figure 1 that the deep level gold mining sector is one of the major contributors to the reduction of demand on the interconnected system over weekends and that

municipal loads are primarily responsible for the formation of "natural" morning and evening peaks and the Winter weekday load swing of 6 GW [3].

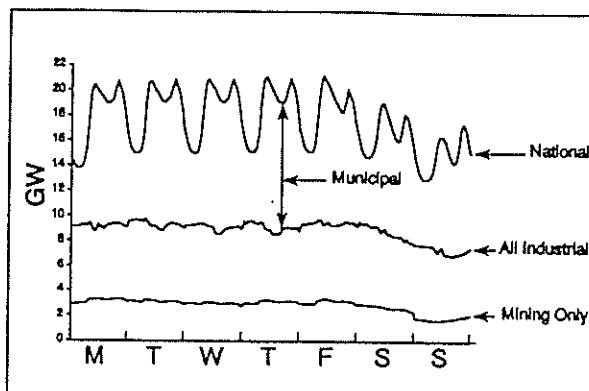


Figure 1: Contribution of deep level mines and other sectors to the total weekly winter demand of South Africa (1989).

The primary aim of this research is to develop a methodology for proving load management potential in existing mines and for least-cost energy system planning of new mines.

Already, as early as 1986, least-cost energy plans were being divided into two main categories [4] namely:

- Minimisation of energy wastage and
- Energy consumption efficiency improvements.

To be able to evaluate the success of any least-cost energy plan, specific energy norms must be set. Analysing the energy norms of existing mines, can be very informative and helpful in determining different energy norms.

2 ENERGY NORMS AND LOAD AUDITS

Energy norms for a few mines are available. These must be collected from the experts working on the mine through an electrical mine audit. The knowledge gained can be used to calibrate the performances of other mines. By doing load audits the mine management can gain valuable insight into the mine's energy cost-effectiveness.

The importance for the existence and use of energy norms can not be emphasized enough. For example, what is the norm for the temperature at the stopes in a deep level mine. Must it be cool enough? What is cool enough? Must another fridge plant be used to make the working temperature in the stopes cool if

nobody is working there? Together with environmental engineering specialists a maximum and minimum working temperature in the stopes should be decided on. The need to operate a cooling plant can then easily follow the same sequence as depicted in figure 3.

Two different types of electrical load audits were identified:

□ "Walk" load audit.

This is a quick load audit, done by a person(s) walking around on a mine looking at procedures and asking questions about the sequence of events on the mine. These inquiries gain valuable information and by analysis give insight into the use of electricity on the mine. This is a labour intensive and time consuming way of doing a load audit and may be subject to the partiality or impartiality of the person doing the audit. An advantage of this "personal" audit is that of not spending a lot of money on measuring equipment.

□ "Measuring" load audit.

For this type of load audit, reliable measuring equipment, either on temporary basis or permanent installation is required and may be costly. A decision must be made on the extensiveness of the data logging, by analysing the detail of the disaggregated load profiles required. If the load is disaggregated into more end-use components of load, then a more reliable audit can be done, at a more expensive price. This type of audit is computer intensive as it takes a lot of computer time to process the measured data. Techniques used to get the insight into the system's performance, shortcomings and positive points exist [5], [6].

To be able to use energy norms and other information gathered with load audits to its full, it is important to be able to have a basic model of the system [7], [8], [9].

3 THE MODELLING STRATEGY

The modelling strategy can be divided into two parts, the control model (top-down) and the end-user group model (bottom-up) as can be seen by the modelling hierarchy in figure 2.

3.1 THE CONTROL MODEL

The conceptual control model block diagram can be seen in figure 3. The control model can be divided into two functions, the calibration function for 'new' mines and the fine tuning function for existing 'instrumented' mines. This is necessary to be able to simulate a new mine with no load profiles available (the calibration function) and to use existing load

profiles to optimise the use of electricity (fine tuning function).

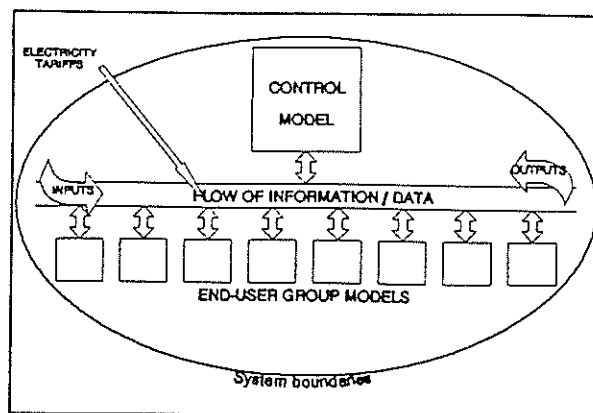


Figure 2 Modelling hierarchy.

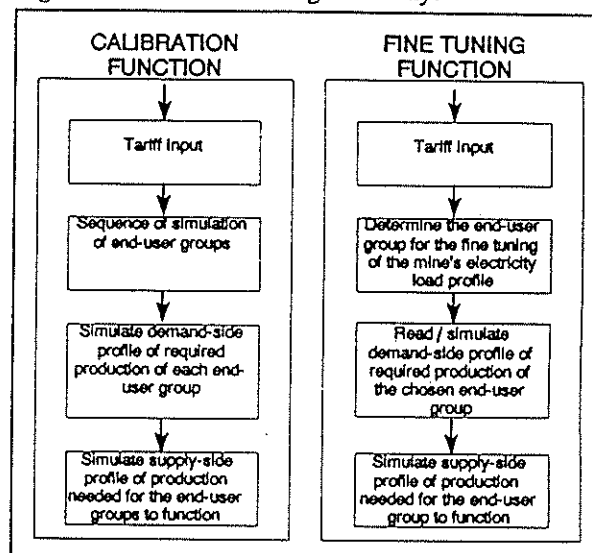


Figure 3 Conceptual control model block diagram.

An example to use the calibration function:

The model can start with a target ore requirement for the mineral processing plant, which chains off into targeted productions for all the other end-user groups. This model must provide, via simulation, the targeted production for each end-user group model and the statistical failing of the end-user group. An example of the influence that a blocked ore pass (not an electricity consumer) can have on the following end-user group, can be seen in figure 4. The actual production of the end-user group can be simulated by the end-user group model. At the end, the actual ore production for the mineral processing plant can be simulated. The electricity demand is derived by linking production or utility volumes to the appropriate energy norms.

An example to use the fine tuning function:

This is possible if data already exists, either by simulation or by measurement. The end-user group's load profile (eg: the compressed air) is then subtracted from that of the whole mine. The fine tuning function

then simulates the optimum machine schedules for the compressors, taking the end-user group configuration (storage, etc.), the electricity tariff, the targeted production and the load profile of the rest of the mine into account. This was already tested [10].

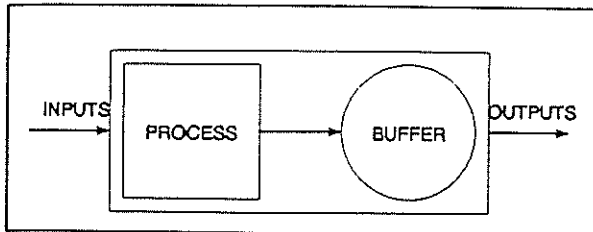


Figure 4 A building block.

3.2 THE END-USER GROUP MODEL

In the research the total mine load profile is disaggregated into eight main end-user groups, i.e. compressors, underground mining, underground pumping, mineral processing plant, hoists, ventilation, fridge plant and hostels and essential services.

A generic end-user group model is being developed, by using a building block concept. A building block, consisting of a process and a storage buffer, see figure 5, can be used to construct an end-user group model. Figure 6 illustrates this building block concept for a simple underground water pumping end-user group.

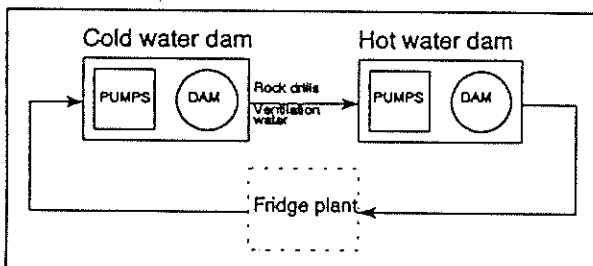


Figure 5 Building blocks for a simple underground water pumping system.

The typical inputs and outputs of each building block are given in figure 7. The concept used for developing the end-user group model is shown in figure 8. The model must first satisfy the system and process limitations and constraints, eg: if the underground hot water dam is going to exceed its maximum water level in the next time interval, the pumps must work. If the pumps need not pump because of limitations and constraints, the model must comply with management requirements, eg: is it possible for something to run during the night or must it work during daytime? If both the previous rules do not make, for example the pumping of water compulsory, the model will take the electricity tariff, the load curves, the end-user group configuration and limitations into account to utilise electricity cost advantages.

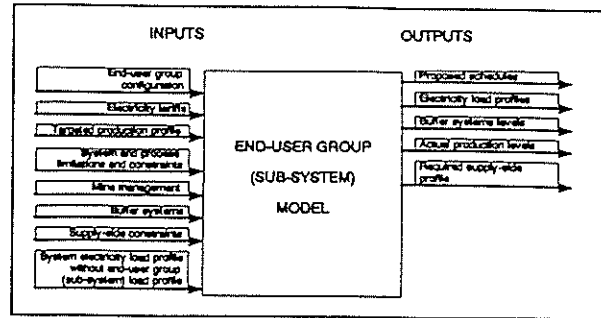


Figure 6 Inputs and outputs for the end-user group model.

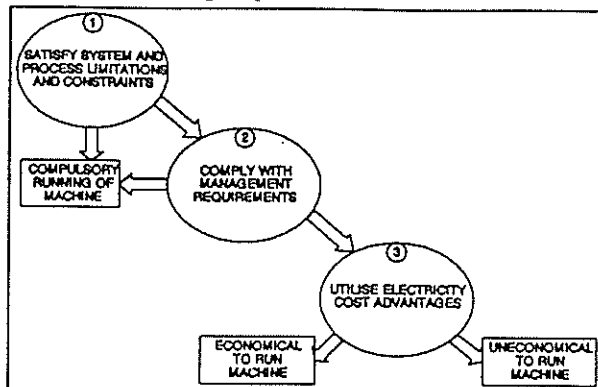


Figure 7 Machine control design logic.

It is very important to note that the driving force behind the building blocks is the flow of the production. This may be the flow of compressed air in m^3/s for the compressed air model, the flow of water in l/s for the underground pumping model or the flow of ore in tons/hour for the mining, hoisting and mineral processing plant models. These flow quantities are converted into electrical quantities by applying the calculated/measured energy conversion parameters for the mine.

4 INTERIM RESULTS

4.1 MINE END-USE SURVEY SYSTEM

Students of the University of Pretoria were assisting with the development of questionnaires for "walk" audits for different end-user groups on a mine, as part of their practical work for a final year subject Energy Management. They were spending a lot of time at a specific deep level mine, asking questions and gaining valuable information from the experts on the mine.

During the preliminary audit at a specific mine, it was discovered that new innovative ways of using electricity more cost-effectively were already implemented by mine personnel, illustrating the importance of expertise of experienced personnel.

4.2 ELECTRICITY COST SIMULATION SYSTEM

Basic model development and testing thereof, using the simulation package SIMAN, for four end-user groups is currently being done at the University of Pretoria. 4 final year students are involved, one on each of the following:

- ☐ compressed air
- ☐ hoists
- ☐ underground pumping and
- ☐ mineral processing.

The modelling philosophy reported in this paper is the result of a major literature study and conceptualisation effort. This forms the major part of our results so far. All current detailed modelling efforts fall within this framework.

4.3 QUICK AUDIT REPORTING SYSTEM

Our first attempt at the report format on the basis of the primary eight end-use categories on a deep-level mine should become available by the end of May 1993. We have already met and interviewed several experts and have started to compile a data base with their details.

5 CONCLUSIONS

The utility companies have to realise that the mining industry probably has the biggest load management potential in South Africa and that expertise already exists. Co-operation and collaboration between mine personnel utility planners and researchers is of the utmost importance for the development of realistic load management techniques for the mining sector.

If the models exist, they would help mine management to understand how expensive unexpected events and inappropriate mining designs can be from an electricity cost point of view. We are optimistic that this project will pave the way for a more formal approach to integrated resource planning with the mining sector. Demand response modelling also promises to indicate the demand-side management potential in deep-level mines.

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7 ACKNOWLEDGEMENTS

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AN END-USER GROUP LOAD MODEL FOR MINERAL PROCESSING ON DEEP LEVEL MINES

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ABSTRACT

The mineral processing plant is an important part of the basic conceptual hierarchical structured energy systems planning model [1] that is being developed to optimise the energy consumption of deep level mines in South Africa.

The mineral processing plant model must accommodate the following factors:

- external influences on the system, such as:
 - * electricity tariffs and industrial legislation.
- internal influences on the system, such as:
 - * system and process limitations,
 - * maintenance and managerial requirements,
 - * production requirements.

1 INTRODUCTION

The mineral processing plant utilises about 20% of the electrical energy consumed by deep level gold mines [2]. An improved management program targeted to improve plant energy consumption offers significant opportunities for modification of the demand profile of a mine. The interconnected system load profile in South Africa could change if most mines follow such a program.

The primary aim of this paper is to discuss the bottom-up approach (for the mineral processing plant end-user group) of the modelling strategy. The differences between this approach and the control model (top-down method) will also be highlighted.

To be able to evaluate the success of the least-cost energy plan for the mineral processing plant, specific energy norms must be determined.

2 ENERGY NORMS AND LOAD AUDITS

There are a few energy norms available for mineral processing plants from the "walk" and "measure" audits that were done on mines [1]. These differ from mine to mine and from plant to plant because of the differences in technology and equipment used.

In many cases there are more energy efficient ways of processing gold bearing ore (compare run-of-mine mills with the conventional washing, crushing and milling), but they involve new and expensive technology. A conversion from old to new technology might prove to be too expensive [3]. It is therefore important to compare the same technology and equipment when energy norms are determined for a specific plant.

To be able to use these energy norms and the other information gathered with the load audits, it is important to be able to have a basic model of the system [3],[4], and understand the role of each end-user group model within this basic model.

3 THE MODELLING STRATEGY

The modelling strategy can be divided into two parts, the control model (top-down) and the end-user group model (bottom-up) as can be seen by the modelling hierarchy in figure 1.

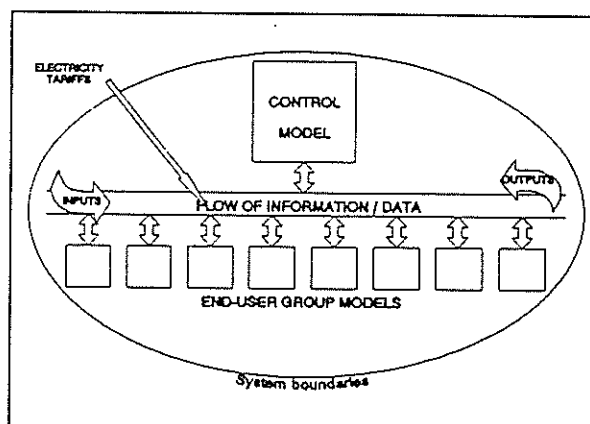


Figure 1 Modelling hierarchy.

3.1 THE END-USER GROUP MODEL

The model developed for the mineral processing plant can provide, via simulation, the necessary outputs required to make the right decisions when deciding on an energy management policy.

A building block concept was used to develop this end-user group model. Each building block consists of a process and a storage buffer [1]. The bottom-up modelling strategy involved the combination of all the building blocks. A full description of their internal dependency (ore movement) was also

necessary for simulation purposes. A building block diagram for the mineral processing plant is shown in figure 2.

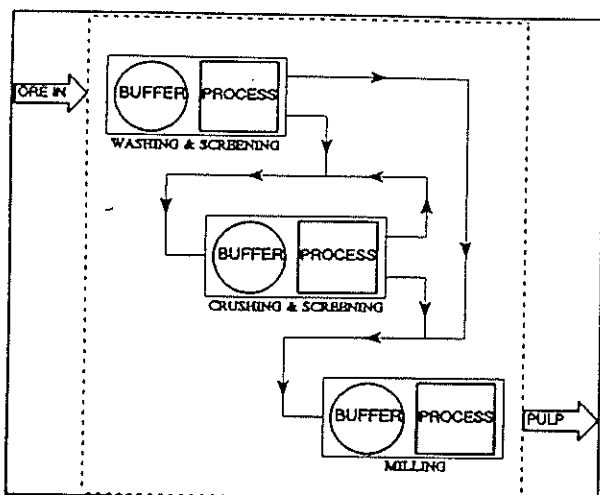


Figure 2 Building block diagram for the mineral processing plant.

This diagram describes the flow of ore through the different processes, and was used to develop the simulation model. The processing of pulp is not included in the simulation model because the continuous processes involved (leaching and CIP), can be considered as being part of the base load.

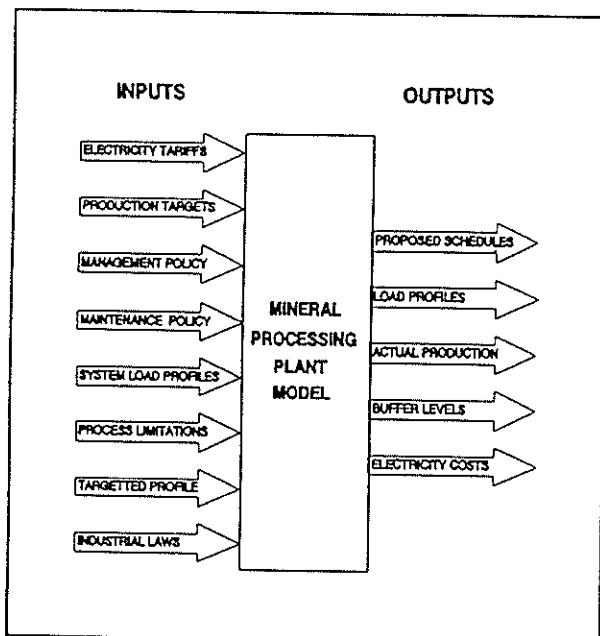


Figure 3 Inputs and outputs for the mineral processing plant building blocks.

The inputs and outputs of each building block for the mineral processing plant are shown in figure 3. These inputs and outputs may differ from the one end-user group to another, and part of the bottom-up approach was to determine exactly what inputs are

needed to get the best (realistic) results or outputs for the end-user group.

3.2 END-USER GROUP CONTROL LOGIC

One of the important results of the initial investigation was the development of a shedding priority list (SPL) for each process on the plant. The SPL for the mineral processing plant is shown in figure 4. Each end-user group will have different shedding priorities, and the importance of each shedding decision may also vary from one group to the other.

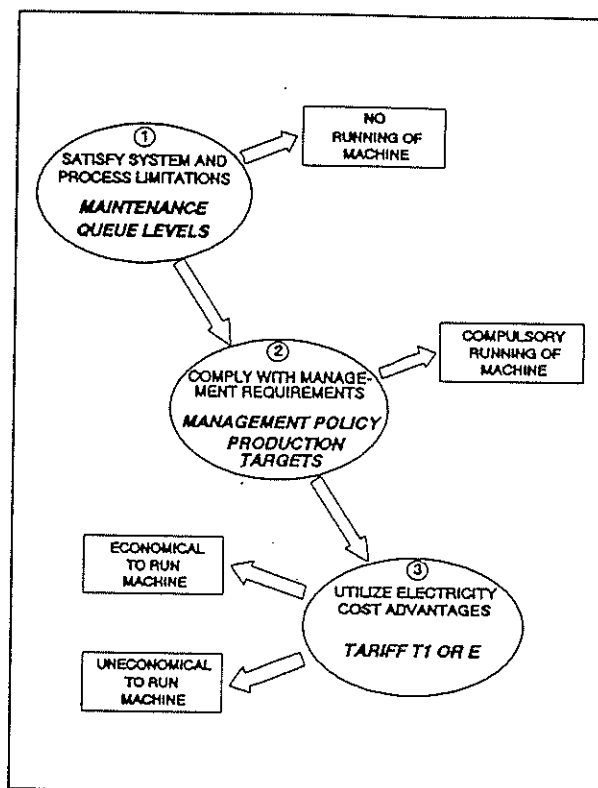


Figure 4 Shedding priority list for the mineral processing plant.

The model must first satisfy system and process limitations, i.e. preventive maintenance requirements and the influence of the availability of ore (buffer levels) at a specific time:

- In order to improve the reliability of the system, preventive maintenance is done almost daily. Management usually determines the time and duration of maintenance according to the availability of manpower and acceptable maintenance standards for each machine.

Preventive maintenance has the highest priority in the SPL, since the production losses inflicted by broken equipment will normally exceed the savings possible when

no maintenance is done and the model only utilises electricity cost advantages.

- When the buffers (bins) feeding a specific process are empty the process must be stopped until ore is available.

If the above conditions for shedding of the loads don't apply, the system must comply with requirements set by management:

- Production targets must be met even if it means that a machine must run during peak (maximum demand control) or expensive (time-of-use control) time. Loads elsewhere on the mine (pumps, winders) must rather be shed to control electricity costs. This implies that other end-user groups must be able to respond to the compulsory load changes that take place on the plant. Costs due to the loss of production will once again exceed electricity costs.

Lastly, the model will take the electricity tariff, load curves and end-user group production limitations into account to utilise possible electricity cost advantages.

4 SIMULATION RESULTS

The most important outputs of the simulation are the plant load profiles. A typical load profile for the plant is shown in figure 5. Production targets were set to maximum values, causing the compulsory running of machines to meet production requirements. The shedding of machines due to maintenance can also be seen on the same graph.

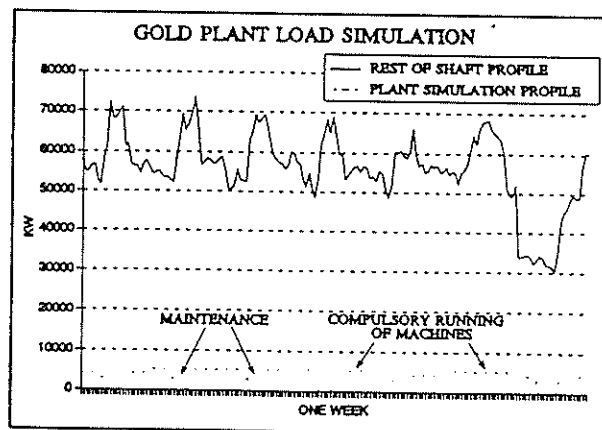


Figure 5 Maximum production plant load curve.

4.1 MAXIMUM DEMAND CONTROL SIMULATION

When production targets are lowered, the compulsory running of machines during expensive or peak time to meet production targets normally becomes unnecessary. The simulation now looks at the tariff structure to utilise electricity cost advantages. The load profile for a maximum demand controlled simulation is shown in figure 6.

It is clear from the figure that the simulation responded to peaks in the load curve of the rest of the shaft (*shaft - no plant*) by shedding loads. This had a significant impact on the value of the peaks reached during maximum demand time. The total shaft load curve was almost 5.1 MW lower during peak times. The effect of Sunday-shutdowns on the shaft total load curve and the drop in the plant load due to maintenance on Sundays can be seen in figure 6, (last day in the week).

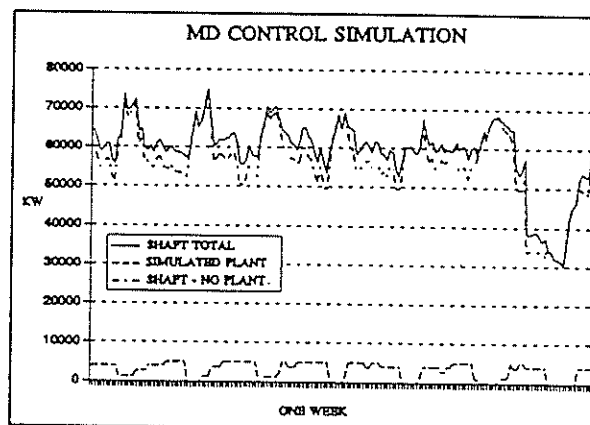


Figure 6 Maximum demand control simulation.

4.2 TIME-OF-USE CONTROL SIMULATION

The plant load profile for a simulation where the loads were controlled according to a time-of-use tariff is shown in figure 7.

The plant simulation responded by shedding as much load as possible to reduce energy costs during the peak times as indicated. If production targets weren't fully met, the control logic will ignore the tariff structure and cause the compulsory running of machines. This can be seen in figure 7, where some loads were still operational during peak time (day 1 and day 3).

Maintenance (both time and duration) has an important influence on the load curve when an attempt is made to utilise time-of-use electricity cost advantages. This can be seen in figure 8 where maintenance was done during off-peak time.

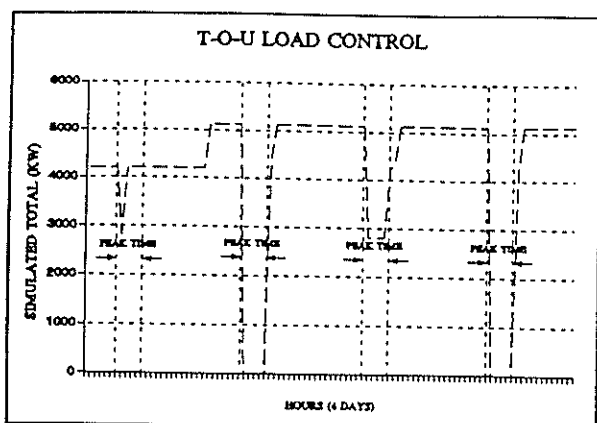


Figure 7 Time-of-use tariff control load profile.

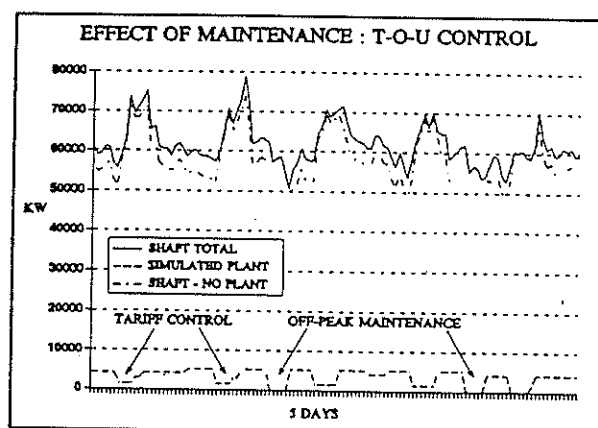


Figure 8 Effect of maintenance - Time-of-use control.

Some loads were shed during peak time, but others had to run to meet production targets, increasing the shaft total load during peak time. If maintenance is rescheduled to peak times the plant immediately utilises the cost advantages in a time-of-use electricity tariff.

5 CONCLUSIONS

This end-user group simulation clearly indicates the existing potential and need for a better demand-side energy management policy on a mineral processing plant. As this and other end-user group simulations become available, it could help mine management to understand how expensive unexpected and unplanned events and even inappropriate management decisions due to a lack of knowledge can be from an electricity cost point of view.

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7 ACKNOWLEDGEMENTS

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NEURAL NETWORKS AS A POWER SYSTEM STATE ESTIMATION TOOL

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State estimation is an important element of an energy management system (EMS). This paper presents a neural computing approach for solving state estimation problem. Its main contribution consists in developing neural network based estimation technique which enables improvements in speed and accuracy of the solution. Testing of the new techniques were performed on examples of state estimation in a small size power system.

1. Introduction

The role of the state estimation is to continuously supply the real-time database of the EMS [1]. At present, the state estimation program is cyclically run every few minutes and/or is triggered by any change of the network topology. Filtered real-time data available as the output of the state estimator is not very useful to the operator because it is not updated fast enough. There is a need for faster state estimator, running at the same speed with the data acquisition process. In such a case the operator can use the state estimated data instead of the raw measurements from SCADA. Application of the neural network technology is one possible way how to improve state estimation.

In this paper we propose the application of a dynamic state feedback neural network the so called Hopfield network [2], in solving the power system state estimation problem. The Hopfield network evolves in time to asymptotically stable equilibrium points, and can be applied in solving optimization problems. The mathematical model of such a network can be represented by the first order system of differential equations [2]:

$$\frac{d\mathbf{X}}{dt} = -\mu \nabla_{\mathbf{X}} \Psi(\mathbf{X}) \quad (1)$$

where μ is a non-negative constant, $\mathbf{X}$ is a vector of problem variables, and Ψ is a objective function of a unconstrained optimization problem. The same neural network could be used for optimization problems with constraints; in these cases the objective function should be reformulated. Special-purpose hardware enables to obtain the solution of large optimization problems in real time; such hardware comprises digital signal processors, array processors, and analog neural processors [3,4].

In this paper the power system state estimation problem is firstly formulated as an unconstrained optimization problem. This is followed by an constrained optimization formulation. For

both problem formulations a Hopfield network solution technique is applied. The results, obtained through simulations, show the advantages of applying the new technique in state estimation.

2. The neural network state estimator

The state of an electrical power network is defined by an $N=n-1$ vector $\mathbf{X}$ of bus voltage phase angles (where n is the total number of nodes). There are M measurements of the active power flows and injections represented by the measurement vector $\mathbf{Z}$. The measurement vector $\mathbf{Z}$ is related to the state vector $\mathbf{X}$ through the linear equation [1]:

$$\mathbf{Z} = \mathbf{H}\mathbf{X} + \mathbf{v} \quad (2)$$

where $\mathbf{H}$ is a constant matrix. The components of the error vector $\mathbf{v}$ are normally distributed random numbers, with zero mean. Moreover, it is assumed that the errors are uncorrelated, and their variances are equal and known.

2.1 Example

The power network model under consideration consists of three nodes (generating, load, and zero injection nodes), and it is shown in Fig.1. The susceptances of the lines are $B_{12}=B_{23}=1\text{pu}$ and $B_{13}=1.5\text{pu}$. The problem is to estimate two phase angles X_1 and X_2 (X_3 is considered zero for the reference node 3) from a set of sampled active power measurements. We chose 4 measurements disposed in the way represented in Fig.1. The constant matrix of linear measurement model (2) is:

$$\mathbf{H} = \begin{bmatrix} 1 & -1.5 & 0 & -1.5 \\ -1 & 0 & -1 & -1 \end{bmatrix}^T \quad (3)$$

The standard deviation of measurements errors is considered to be 0.1pu .

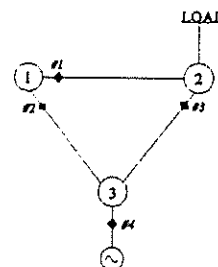


Figure 1 3-bus example

2.2 The unconstrained formulation

The measurements are sampled over constant length intervals (T), and the instantaneous

square error function for the interval $kT \leq t < kT+T$ is:

$$\Psi_k(\mathbf{X}, t) = \quad (4)$$

$$= [\mathbf{Z}^{(k)} - \mathbf{H}\mathbf{X}(t)]^T [\mathbf{Z}^{(k)} - \mathbf{H}\mathbf{X}(t)]$$

The gradient of the above function is:

$$\nabla_{\mathbf{X}} \Psi_k(\mathbf{X}, t) = -2\mathbf{H}^T \mathbf{Z}^{(k)} + 2\mathbf{H}^T \mathbf{H}\mathbf{X}(t) = \quad (5)$$

$$= -2 \sum_{i=1}^M \varepsilon_i(t, k) \mathbf{H}_i$$

where:

$$\varepsilon_i(t, k) = \mathbf{Z}_i^{(k)} - \mathbf{H}_i \mathbf{X}(t) \quad (6)$$

$$\mathbf{H} = [\mathbf{H}_1^T \mathbf{H}_2^T \dots \mathbf{H}_i^T \dots \mathbf{H}_M^T]^T$$

According to the neural network optimization approach (1) the state of the power system would correspond to the solution of the system of differential equations:

$$\frac{d\mathbf{X}(t)}{dt} = 2\mu \sum_{i=1}^M \varepsilon_i(t, k) \mathbf{H}_i \quad (7)$$

The block diagram of a neural estimator as a direct implementation of the system of differential equations (7) is shown in Fig.2.

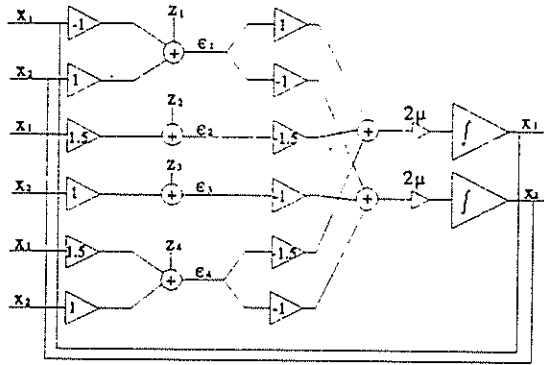


Figure 2 An unconstrained neural estimator for a 3-bus network

2.3 The constrained formulation

The zero injections at a switching stations are treated as virtual measurements. Such measurements are incorporated in the state estimation formulation as an equality constrain $\mathbf{A}\mathbf{X} = \mathbf{0}$, where $\mathbf{A}$ is a constant matrix and $\mathbf{X}$ represents the state vector. For the three-bus example the switching substation is situated at node 1. The constraint equation, corresponding to a zero injection at node 1, is represented in matrix form as $[2.5 \ -1][\mathbf{X}_1 \ \mathbf{X}_2]^T = \mathbf{0}$. The standard method of solving a constrained estimation problem consists in using Lagrange multipliers [5].

For the neural network estimation approach, a variable-reduction technique [5] has been investigated. This method transforms a constrained problem into an unconstrained one, and reduces the number of independent variables. Let use the following partition of the constraint matrix $\mathbf{A}$ and of the state vector $\mathbf{X}$:

$$[\mathbf{U} \ \mathbf{V}] \begin{bmatrix} \mathbf{X}_v \\ \mathbf{X}_u \end{bmatrix} = \mathbf{0} \quad (8)$$

It follows that the relation between $\mathbf{X}_v$ (dependent vector of p variables) and $\mathbf{X}_u$ (independent vector of $N-p$ variables) is:

$$\mathbf{X}_v = -\mathbf{V}^{-1} \mathbf{U} \mathbf{X}_u \quad (9)$$

where $\mathbf{V}$ is $p \times p$ non-singular matrix. For the state vector $\mathbf{X} = \mathbf{C}\mathbf{X}_u$ (where $\mathbf{C} = [-\mathbf{V}^{-1}\mathbf{U} \ \mathbf{I}]^T$, and $\mathbf{I}$ is identity matrix) the reduced gradient of the instantaneous square error function (4) can be derived:

$$\nabla_{\mathbf{X}} \Psi_k(\mathbf{X}_u, t) = \frac{\partial \Psi_k(\mathbf{X}, t)}{\partial \mathbf{X}} \frac{\partial \mathbf{X}}{\partial \mathbf{X}_u} = \quad (10)$$

$$= \mathbf{C}^T \nabla_{\mathbf{X}} \Psi_k(\mathbf{X}, t) = -2 \sum_{i=1}^{M-p} \varepsilon_i(t, k) \overline{\mathbf{H}}_i$$

where

$$\overline{\varepsilon}_i(t, k) = \mathbf{Z}_i^{(k)} - \overline{\mathbf{H}}_i \mathbf{X}_u(t) \quad (11)$$

$$\overline{\mathbf{H}}_i = \mathbf{H}_i \mathbf{C}$$

A neural estimator with linear equality constraints and using a variable-reduction technique that comprises equation (9) and reduced system of differential equations (12) for a three bus network is represented in Fig.3.

$$\frac{d\mathbf{X}_u(t)}{dt} = 2\mu \sum_{i=1}^{M-p} \overline{\varepsilon}_i(t, k) \overline{\mathbf{H}}_i \quad (12)$$

The advantage of using a variable-reduction approach compared to using the Lagrange multipliers consists in a simpler electrical circuit realization and in a faster solution.

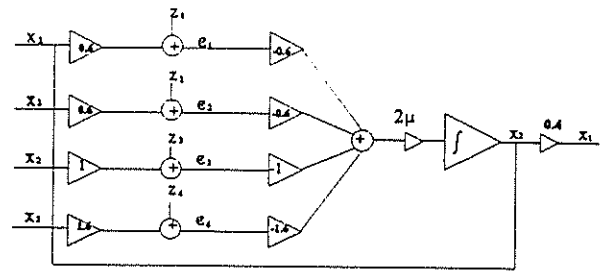


Figure 3 A constrained neural estimator for a 3-bus network

3. Simulation results

In this section we present results of computer simulation of neural network estimation techniques. For a 3-bus power system example, two neural networks were developed: an unconstrained estimator (Fig.2) and a constrained estimator (Fig.3). Simulations were performed by solving a system of differential equations: the system (7) for the unconstrained estimator and the system (12) for the

constrained estimator.

In the first case we tested an unconstrained estimator under the assumption that the measurements were constant signals corrupted with errors (see section 2.1). Expected values of measurements were $Z_1=Z_2=0.375$ pu, $Z_3=0.625$ pu, and $Z_4=1$ pu. The state estimated values were $X_1=-0.25$ rad, and $X_2=-0.625$ rad. In the simulation study, the measurements were sampled over the time interval $T=1$ s. At the beginning of the simulation, the states are unknown; therefore starting values for state variables were randomly chosen. Results of simulation are presented on the fixed contour plot of the quadratic performance surface:

$$\Psi(X) = E[Z_k^T Z_k] - 2E[Z_k^T] HX + X^T H^T HX \\ = 5.5X_1^2 + 3X_2^2 + X_1X_2 + 3.375X_1 + 4X_2 + 1.67$$

Figures 4 and 5 presents results of the unconstrained neural estimator simulation for different gain constants. For a gain value of 0.005 the estimator operates slower but the results are more accurate then for a gain value of 0.1.

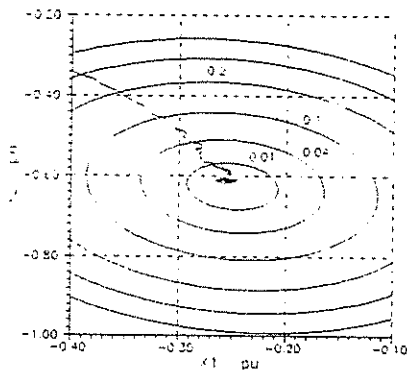


Figure 4 $T = 1$ s, $\mu = 0.005$

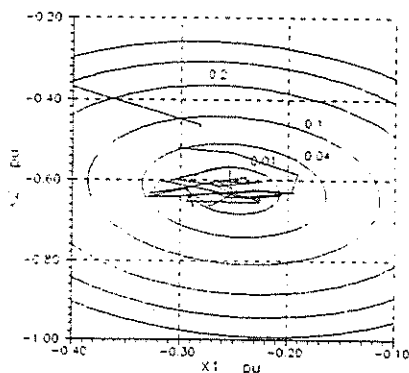


Figure 5 $T = 1$ s, $\mu = 0.1$

The second part of this section presents results of simulation obtained with a neural network tracking of state variables, during variation of load conditions. The performances of the unconstrained and constrained neural estimators are compared. The following performance index PI indicates the

filtering capability of the different estimators:

$$PI(t) = \frac{\sum_{i=1}^M (z_i^E(t) - z_i^T(t))^2}{\sum_{i=1}^M (z_i^M(kT) - z_i^T(t))^2}$$

where z_i^T , z_i^M , and z_i^E are the true, measured and estimated values of the i -th measurement. True and estimated values are changing continually in time and measured values are discrete (constant in each snapshot interval). The numerator of the performance index indicates how close are the estimated values to the true values, and the denominator shows the level of uncertainty of the measurements. In the case of a good filtering, performance index should be always smaller than 1.

The simulation study of unconstrained and constrained WLS estimators, and unconstrained and constrained neural estimators was made over a period of 70s. Load and generation are constant in the first 20s. After that, sudden drop of load (from 1pu to 0.8pu) take place following with slow decreasing of load in next 40s (from 0.8pu to 0.7pu). In the last 10s the operating point of the system is constant again. Representative results of simulation are presented on Figures 6, 7, and 8. Parameters of unconstrained and constrained neural estimators (Figures 2, and 3) are $T=1$ s, and $\mu=0.05$.

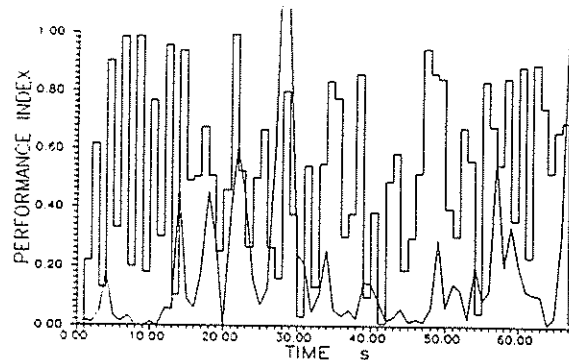


Figure 6 PI's for unconstrained WLS estimator (discrete line) and unconstrained neural estimator (continual line)

Figure 6 shows difference of PI's between classical unconstrained WLS tracking (discrete line) and continual tracking with unconstrained neural net from Fig.2. Comparison between tracking of the state variable X_1 obtained with unconstrained and constrained WLS (discrete lines), and neural approach (unconstrained and constrained versions) are presented at Figure 7. Probability density functions of state values residuals (difference between estimated and true values for X_1 and X_2 for every sample of measurements) are presented at the Fig.8. These curves clearly indicate filtering performances of unconstrained and constrained WLS estimators (WLSE) and unconstrained and

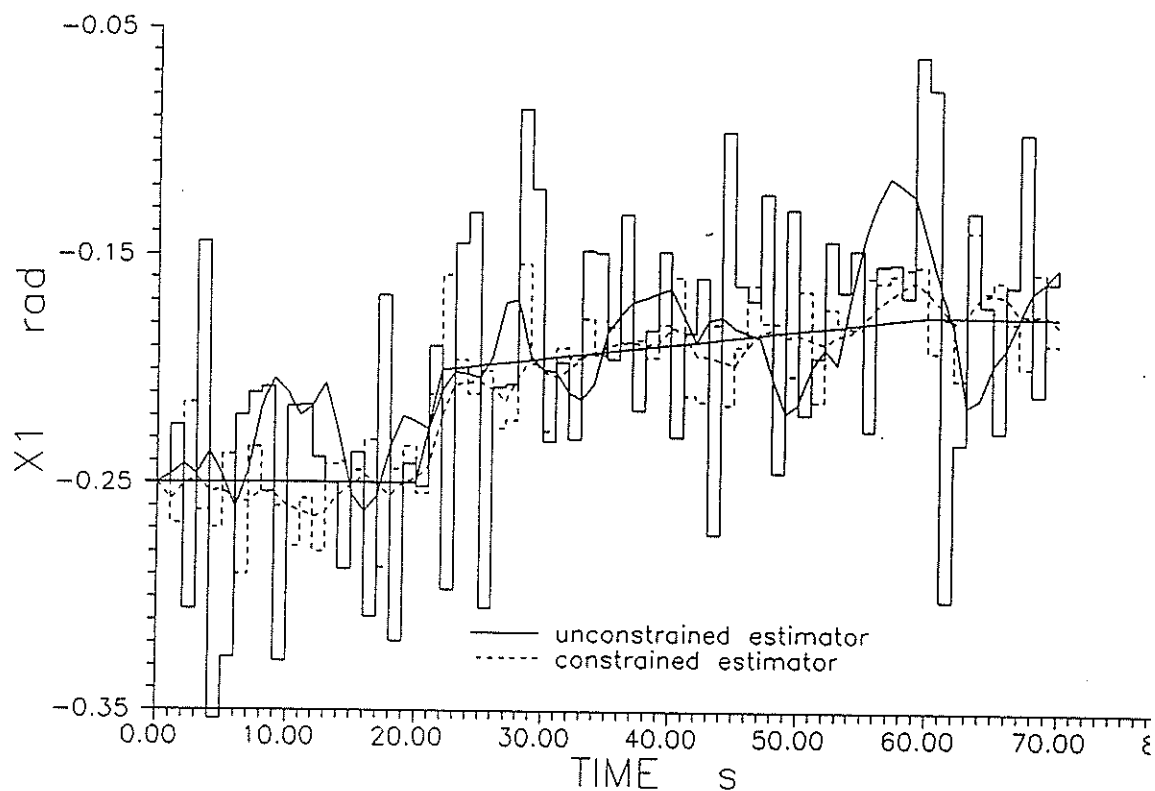


Figure 7 Tracking of the state variable X_1 (3-bus example) with WLS estimators (solid and dashed discrete lines), and neural estimators (solid and dashed continual lines)

constrained neural estimators (NE).

4. Conclusions

From the presented results the following remarks are to be made:

- For the slow varying measurements, the neural estimator with smaller gain constant gives more accurate results. Shorter sampling interval also produces more accurate results.
- For the sudden change of loading conditions the neural estimator with smaller gain constant gives non accurate results (see Fig.6). If the processing of neural estimator is more noise insensitive, then fast adapting capabilities suffer and vice versa.
- The constrained neural estimator produces more accurate results (see Figures 6,7 and 8) and makes neural circuit simpler (Figure 3).
- The neural estimator (both unconstrained and constrained) produces better results (in terms of accuracy and speed) as compared with the WLS classical state estimator (see Figures 6, 7, and 8).

The presented new power system state estimation technique was verified only for a small three-bus network and for the case of a linear or DC-load flow approximation. Preliminary results show that extension to large networks and AC-load flow is feasible with the new neural network approach and that improvements in both accuracy and speed could be achieved in the future power

system control centres.

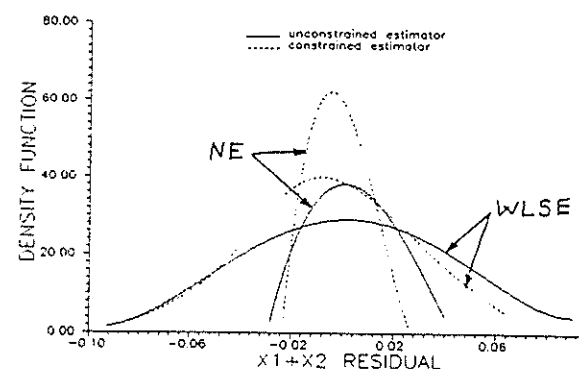


Figure 8 Probability density functions for WLS estimators (WLSE) and neural estimators (NE)

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IMPLEMENTATION OF A NEURAL NETWORK INDUCTION MACHINE CONTROL STRATEGY

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ABSTRACT

This paper briefly presents an investigation into problems and possible improvements that were highlighted during the development, by the second and third authors, of an adaptive Artificial Neural Network (ANN) induction motor identification and control strategy [1]. The objective of this investigation was to evaluate the proposed system in terms of practical hardware implementation implications and make any appropriate modifications before moving on to the synthesis phase of this project. The sigmoidal feedforward ANN used and its advantages over Recursive and Radial Basis Function ANN's are presented. The backpropagation training algorithm is used and its advantages over other algorithms are also described.

1. INTRODUCTION

The most common system identification and control applications of ANN's involve the off-line training of arbitrarily large (many hidden nodes and many nodes in each layer) feedforward sigmoidal ANN's using the backpropagation training algorithm [1]. This training is done using large amounts of data taken directly from the process to be identified and controlled and must include fairly dense data points on the extreme boundaries of system operation. Comparatively sparse data points may be used to represent system behaviour within these boundaries as ANN's have inherently good generalisation (interpolation) capabilities but generally exhibit very poor performance outside (extrapolation) of the operating conditions over which they are trained. Disadvantages of this approach are that the amount of system data, training time and ANN size are not only dependent on system complexity but also the range of operating conditions involved. These systems are also prone to degraded performance or failure should unforeseen circumstances force the system out of the range of operating conditions over which the ANN has been trained.

One of the most attractive features of the proposed system is that it remains adaptive in nature throughout operation, which gives it the ability to track parameter variations under all operating conditions. This is

achieved by continually updating the weights of the ANN controllers in accordance with on line weight modifications effected in ANN system identifiers by the training algorithm. Since training is done on-line, it is imperative that the ANN size and execution time of the ANN output calculation and training algorithm be minimised in order to allow implementation on modest hardware while maintaining the desired sampling rate.

2. COMPARISON OF ANN TYPES

(a) Sigmoidal Feedforward ANN's

These ANN's are characterized by a layered structure of elements called nodes, as shown in figure 1.

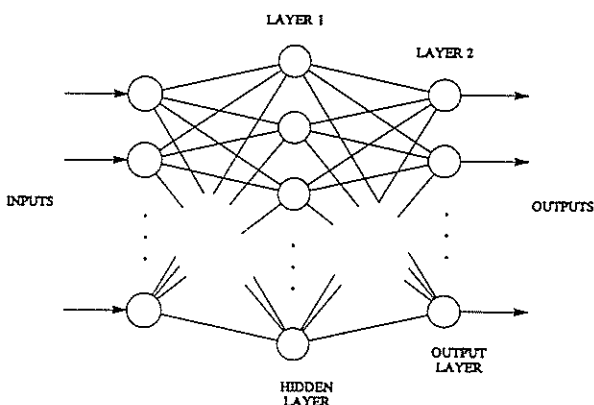


Figure 1: A Two Layered Feedforward ANN

The output of each node in a given layer is connected via weights to the input of every node in the subsequent layer. The layers to the left of the output layer are called hidden layers, and consist of nodes called perceptrons, as shown in fig. 2.

The first of two distinct modes of operation is the feedforward mode, where inputs are applied to the ANN and propagated through it to the outputs. The other mode of operation is the training mode, where the ANN weights are modified according to a training algorithm aimed at minimizing the error between the actual ANN output and the desired output, which is some (unknown) function of the input variables.

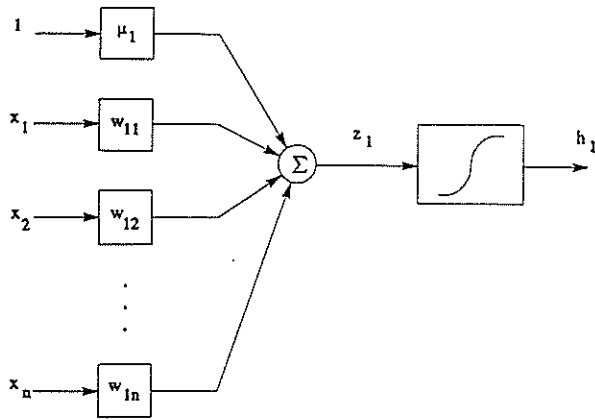


Figure 2: The Sigmoidal ANN Perceptron

One advantage of the sigmoidal ANN is that it has formally been shown [2] that only one hidden layer (with sufficient nodes) is required in order to adequately approximate a given function over a specified range. The proposed scheme exploits this property and uses sigmoidal ANN's with one hidden layer exclusively and determines the number of hidden nodes required via a heuristic approach. Further advantages of this type of ANN become evident after an investigation of the backpropagation training algorithm in section 3.

Consider a sigmoidal feedforward ANN with n inputs, r outputs and one hidden layer of m hidden nodes. The activation of the j th node in the hidden layer z_j is a weighted sum of its inputs x_i and threshold μ_j .

$$z_j = \sum_{i=1}^n w_{ji} x_i + \mu_j \quad (1)$$

The activation of the j th hidden node z_j is fed through the sigmoidal function to obtain its output h_j .

$$h_j = a(z_j) = (1 + e^{-z_j}) \quad (2)$$

The k th output y_k is simply a weighted sum of the hidden layer outputs h_j .

$$y_k = \sum_{j=1}^m w_{kj} h_j \quad (3)$$

where: $W1 \in \mathbb{R}^{m \times n}$ is the input weight matrix
 $W2 \in \mathbb{R}^{r \times m}$ is the output weight matrix
 $\underline{x} \in \mathbb{R}^n$ is the input vector
 $\underline{\mu} \in \mathbb{R}^m$ is the threshold vector
 $\underline{z} \in \mathbb{R}^m$ is the activation vector

$\underline{h} \in \mathbb{R}^m$ is the hidden layer output vector
 $\underline{y} \in \mathbb{R}^r$ is the output vector

The y_k are not passed through a sigmoidal function since outputs in the proposed system are required to attain real values which may lie outside of the sigmoidal output interval of (0,1).

It is convenient to consider the thresholds μ_j as additional input weights connecting a unit input to each of the hidden nodes. This is done by augmenting the input vector $\underline{x}$ with a unit element, i.e. $\underline{x} = [1 \ x_1 \ x_2 \dots x_n] \in \mathbb{R}^{1+n}$, and augmenting $W1$ by inserting $\underline{\mu}$ as the first column, i.e. $W2' = [\underline{\mu} \ W2] \in \mathbb{R}^{m \times (1+n)}$. The thresholds may thus be updated as weights.

(b) Radial Basis Functions (RBF's)

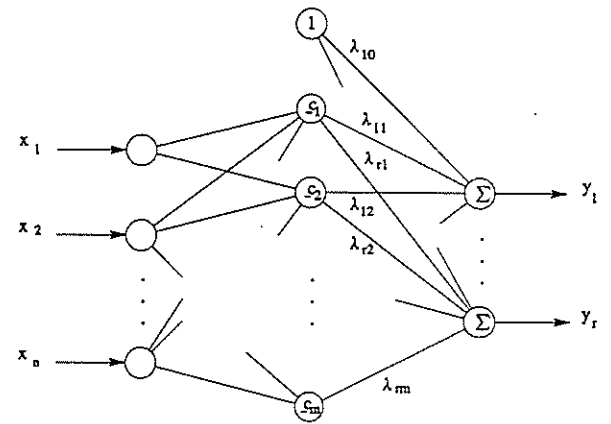


Figure 3: A Radial Basis Function Expansion

RBF's essentially have the same layered feedforward structure as sigmoidal ANN's with one hidden layer. RBF's, however, have only unity input weights associated with them, and each node in the hidden layer is assigned a vector value which is referred to as its centre c_i .

Once again, consider a network with n inputs, r outputs and m hidden nodes. The activation of the j th hidden node z_j is given as the euclidean norm $\|\cdot\|$ of the distance between its centre c_i and the input vector $\underline{x}$.

$$z_j = \|\underline{x} - \underline{c}_j\| \quad (4)$$

One of four main activation functions $\phi(\cdot)$ are used, and the k th output y_k is calculated as a weighted sum of the outputs of the hidden nodes z_j and a threshold

λ_{k0} .

$$y_k = \lambda_{k0} + \sum_{j=1}^m \lambda_{kj} \phi(\|x - c_j\|) \quad (5)$$

The four main forms of function $\phi(\cdot)$ are [3] the *thin-plate-spline* function,

$$\phi(v) = v^2 \log(v) \quad (6)$$

the *gaussian* function,

$$\phi(v) = e^{-\frac{v^2}{\beta^2}} \quad (7)$$

the *multiquadratic* function

$$\phi(v) = \sqrt{v^2 + \beta^2} \quad (9)$$

and the *inverse multiquadratic* function

$$\phi(v) = \frac{1}{\sqrt{v^2 + \beta^2}} \quad (9)$$

where: β is a real constant.

$\lambda_0 \in \mathbb{R}^r$ is the *output threshold vector*

$\lambda \in \mathbb{R}^{r \times m}$ is the *output weight matrix*

The main disadvantage of RBF's is that the method of iteratively updating the centres, known as *orthogonal forward regression* (OFR), is not well suited to an on-line process and must thus be carried out off-line. A formal presentation of this method is beyond the scope of this paper, but may be found in [3], [4]

With fixed centres, the only parameters that may be adaptively modified in RBF's are the output weights λ . Consider a RBF and a sigmoidal ANN of the same size. While there are fewer weight adjustments to be made during on-line training in a RBF, this intuitively makes it more rigid and thus a larger RBF would be needed in order to give comparable tracking performance.

A further disadvantage of RBF's in relation to sigmoidal ANN's is that the feedforward, or output, calculation requires more *floating point operations* (FLOPS) in the case of a RBF. This is largely due to the calculation of the euclidean norm in equation (5). It is assumed that a lookup table is used to evaluate the activation function in each case and thus this operation is computationally identical, irrespective of the function used.

It is argued [3] that RBF's offer superior approximation capabilities to those of sigmoidal ANN's, but this analysis is primarily aimed at off-line applications, where there are no time constraints on the determination of the centres or the overall size of the network, and not adaptive ones where time and size are both limited.

(c) Recurrent ANN's (RANN's)

RANN's are fully connected networks of nodes, as shown in figure 3, and are normally used in conjunction with sigmoidal activation functions.

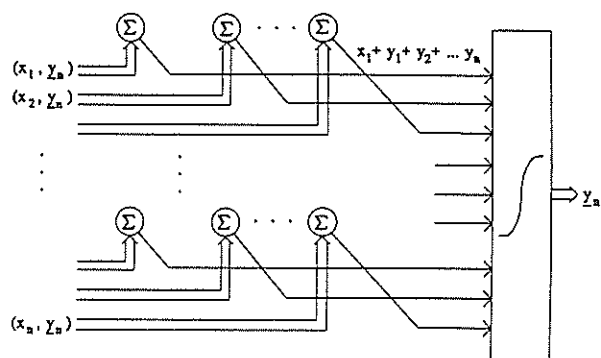


Figure 4: Fully Connected Recurrent ANN (RANN)

It can be seen that the output of each node in the network forms an input to every other node. The advantage of this configuration is that it incorporates a large amount of internal feedback and becomes a dynamic system as opposed to the static sigmoidal ANN's and RBF's. Detailed stability studies based on Lyapunov Analysis have been done on RANN's [5], [6], whereas no parallel has been achieved with either sigmoidal ANN's or RBF's.

An inherent advantage of RANN's is that they may be used directly to model dynamic systems, without any external dynamic elements, since their outputs are not statically determined after one feedforward calculation and a number of iterations are required before the outputs converge.

The major disadvantage of RANN's, however, is that the internal feedback introduces many closed loops and makes them highly serial in nature and thus much more computationally complex and slow than the highly parallel sigmoidal ANN's and RBF's. It can also be seen that the fully connected structure of RANN's incorporates many more weights than a layered structure.

3. BACKPROPAGATION TRAINING

(a) The Backpropagation Algorithm

This is by far the most popular sigmoidal feedforward ANN training algorithm in general use and in particular for system identification and control applications. The reason for its popularity probably lies in the fact that it is relatively simple and possibly the best understood method. Since backpropagation is an error based algorithm, it may be said that the ANN is made to "learn from its mistakes", which is intuitively easy to understand.

It has may be shown [7] that backpropagation is a simplification of many other algorithms, such as the *recursive error prediction algorithm* and so called *second order methods*, which indicates that it is representative of a fundamental approach.

The basis of the on-line algorithm is that each weight in the ANN is adjusted at time $(T+1)$ according to an estimate of its contribution to the corresponding output node error at time (T) , where $T \in \mathbb{I}$, and the unit of time is the sampling rate T_s . If $y_{kd}(T)$ is the k th desired output and $y_k(T)$ the k th ANN output, then the k th output node error $\delta_{yk}(T)$ is given by:

$$\delta_{yk}(T) = y_{kd}(T) - y_k(T) \quad (10)$$

and the j th hidden node activation equivalent error $\delta_{zj}(T)$ is given by:

$$\begin{aligned} \delta_{zj}(T) &= \frac{dh_j(T)}{dz_j(T)} \sum_{k=1}^r \delta_{yk}(T) w_{kj}(T) \\ &= h_j(T)[1 - h_j(T)] \sum_{k=1}^r \delta_{yk}(T) w_{kj}(T) \end{aligned} \quad (11)$$

The changes in output weights Δw_{2kj} are given by:

$$\Delta w_{2kj}(T) = \gamma_m \Delta w_{2kj}(T-1) + \gamma_g \delta_{yk}(T) h_j(T) \quad (12)$$

and changes in input weights Δw_{1ji} are given by:

$$\Delta w_{1ji}(T) = \gamma_m \Delta w_{1ji}(T-1) + \gamma_g \delta_{zj}(T) x_i(T) \quad (13)$$

where: $\gamma_m, \gamma_g \in [0,1]$ are the *momentum* and *gain* constants respectively.

The weights at time $(T+1)$ are simply:

$$\begin{aligned} w_{1ji}(T+1) &= w_{1ji}(T) + \Delta w_{1ji}(T) \\ w_{2kj}(T+1) &= w_{2kj}(T) + \Delta w_{2kj}(T) \end{aligned} \quad (13)$$

Simulations [1] have shown on-line backpropagation training of sigmoidal ANN's to converge rapidly in the proposed scheme and its simplicity makes it the most practically viable one for implementation.

(b) Backpropagation and Sigmoidal ANN's

An advantage of the backpropagation training algorithm used in conjunction with sigmoidal feedforward ANN's is the convenient form the derivative of the sigmoidal function, as shown in equation (11), which eliminates the need for a look up table for this function.

A further advantage of the backpropagation algorithm is that it compliments the highly parallel nature of the sigmoidal feedforward ANN in that weight adjustments are based only on local information. As such, it may be argued that, in the worst case, one microprocessor could be assigned to each node in a given network in order to achieve the desired sampling rate.

4. PROPOSED CONTROL STRATEGY

(a) ANN NARMAX Model Identification

It has been discussed how two layer sigmoidal feedforward ANN's, simply referred to as ANN's from this point onwards, may be trained by backpropagation to identify a given nonlinear vector function or input/output mapping. This ability may be used to identify nonlinear sampled, or discrete-time, systems if the system to be identified may be represented by a nonlinear input/output model or mapping. Most systems may be represented by NARMAX (Nonlinear AutoRegressive Moving Average with eXogenous inputs) model, which is of the desired form. The NARMAX model of a system gives its output y at time $(k+1)$ (again where $k \in \mathbb{I}$ and the unit of time is the sampling rate T_s) as some nonlinear function $f(\cdot)$ of current, time k , and previous, time $(k-d)$, input u and output y as [1], [8]:

$$y(k+1) = f(y(k), \dots, y(k-d), u(k), \dots, u(k-d_u)) \quad (15)$$

where $d_u, d_y \in \mathbb{I}$ are the maximum delays in the input and output vectors u, y respectively.

(b) Induction Machine Stator Current Control

The $\alpha\beta$ stator currents $i_s^{\alpha\beta}$ of the induction machine are given in terms of present and delayed values of $i_s^{\alpha\beta}$, rotor speed ω_r and $\alpha\beta$ stator voltages $v_s^{\alpha\beta}$ are given by [1], [8]:

$$i_s^{\alpha\beta}(k+1) = f[i_s^{\alpha\beta}(k), i_s^{\alpha\beta}(k-1), \omega_r, \omega_r(k-1), v_s^{\alpha\beta}(k-1)] + c_v v_s^{\alpha\beta} \quad (16)$$

An ANN is trained on-line to identify $i_s^{\alpha\beta}(k+1)$ and c_v , which are used to determine the desired control input $v_s^{\alpha\beta}(k)$ from equation (16) as:

$$v_s^{\alpha\beta}(k) = \frac{f(i_s^{\alpha\beta}(k), i_s^{\alpha\beta}(k-1), \omega_r(k), \omega_r(k-1), v_s^{\alpha\beta}(k-1)) + i_s^{\alpha\beta}(k+1)}{c_v} \quad (17)$$

(c) Induction Machine Rotor Speed Control

A similar approach is taken to identify the NARMAX representation of the current fed model of the induction machine. This model is expressed in terms of the change in rotor speed $\Delta \omega_r$ (as opposed to the absolute rotor speed), as a function of present and delayed values of $\Delta \omega_r$, the dq stator currents i_s^{dq} , slip s and electrical supply frequency ω_e as:

$$\Delta \omega_r(k+1) = f_m(\Delta \omega_r(k), \Delta \omega_r(k-1), i_s^{dq}(k), i_s^{dq}(k-1), s\omega_e(k), s\omega_e(k-1), \omega_r(k)) \quad (18)$$

The ANN thus identifies the mechanical load dynamics. Unfortunately, this model is not linear with respect to its inputs ω_e and i_s^{dq} , as in the case of equation (16) and so no control equation such as (17) can be directly applied. In this case, a second ANN is used to control the electrical supply frequency ω_e , and a conventional PI controller is used to set the dq stator current reference. The ANN controller is trained on the backpropagated ANN identifier error since the error in electrical supply frequency cannot be determined directly.

5. CONCLUSION

A brief comparison of different ANN types and the backpropagation training algorithm, as applied to adaptive (on-line) system identification and control, has been presented. The advantages of backpropagation trained two layer sigmoidal feedforward ANN's, with regard to practical implementation, have also been discussed.

From the overview of the induction machine control strategy presented, it can be seen that the electromagnetic model identification and stator current control is optimal. Further investigation is to be done to determine the existence of a similar method for control of the rotor speed, but this is to be done in parallel with the practical implementation of the proposed scheme. At this stage, a 8096's array and transputers are being considered as possible hardware platforms.

The question of position control is to be addressed at a later stage, once a prototype system has been

commissioned.

6. ACKNOWLEDGEMENT

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IDENTIFICATION AND CONTROL OF INDUCTION MACHINES USING ARTIFICIAL NEURAL NETWORKS

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ABSTRACT

This paper proposes the use of Artificial Neural Networks (ANNs) to identify and control the induction machine. Two systems are presented; a system to adaptively control the stator currents via identification of the electrical dynamics, and a system to adaptively control the rotor speed via identification of the mechanical and current-fed system dynamics. Various advantages of these control schemes over other conventional schemes are cited and the combined speed and current control scheme is compared with the standard vector control scheme.

1. INTRODUCTION

The induction machine is a nonlinear system and in addition many of its parameters vary with time and operating condition. Conventional induction machine dynamic control techniques such as vector control and direct-self-control (DSC) attempt to reduce the complex nonlinear dynamic structure into a linear structure, in order to enable the application of linear design techniques. Although the implementation of both methods has been largely successful, they both suffer from sensitivity to parameter variations. For vector control, numerous techniques (eg. [1]) have been proposed to overcome this problem with varying degrees of success. In addition, both conventional techniques are relatively complex and require the use of relatively highly trained commissioning personnel.

Recently, the use of Artificial Neural Networks (ANNs) to identify and control nonlinear dynamic systems has been proposed [2]-[4]. This paper investigates the use of ANNs to identify and adaptively control the induction machine. Previous papers [5],[6] have proposed schemes which teach ANNs off-line to mimic existing stator current controllers. Once sufficiently well-trained, the ANNs replace the original controllers, resulting in certain advantages over the original controllers, such as increased speed of execution and fault tolerance. The schemes proposed in this paper identify the induction machine and then control it in either a direct or indirect on-line manner. They remain adaptive in nature throughout operation.

The paper starts with a brief introduction to ANNs and presents a commonly used learning algorithm. The overall control problem is separated into two subproblems,

namely control of the stator currents and control of the rotor speed.

For control of the stator currents, the conventional electromagnetic model of the induction machine expressed in two-axis stationary coordinates is presented and an observable form of this model is introduced. A system to identify the electromagnetic model using an ANN is introduced and a system to adaptively control the stator currents, based directly on the identification model, is proposed. Simulations are then presented to demonstrate the performance of the proposed system.

For control of the rotor speed, the current-fed model of the induction machine is presented and an observable form of this model is introduced. A system to identify the change in rotor speed using an ANN is introduced and a system to adaptively control the rotor speed using an ANN controller trained indirectly via the identification ANN, is proposed. Simulations are presented to demonstrate the performance of the system and a comparison is made with a standard vector control scheme.

2. ARTIFICIAL NEURAL NETWORKS

Artificial neural networks (ANNs) are computing systems whose structures are inspired by a simplified model of the human brain. Many different types of ANNs have been

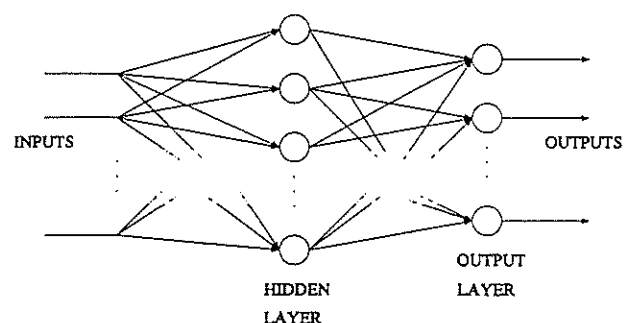


Fig. 1. A Two-Layer ANN

proposed, however this paper considers sigmoidal feedforward ANNs exclusively.

(a) Mathematical Model of Feedforward ANN

Fig. 1 shows a typical two-layer ANN. It consists of two layers of simple processing units (termed *nodes* or *neurons*). Each input is connected via a weight to each (hidden) node in the *hidden layer*. Each hidden node is, in turn, connected via a weight to each (output) node in the *output layer*. The output of hidden node j is given by $h_j = a(z_j)$ where $a(\cdot)$ is termed the activation function and

$$z_j = \sum_i w_{ji} x_i + \mu_j \quad (1)$$

is termed the activation of node j and is the weighted sum of the inputs x_i to node j and the threshold μ_j of node j . For the systems described in this paper, thresholds are unnecessary and are omitted. The activation function used in this paper is the sigmoidal function:

$$h_j = a(z_j) = (1 + e^{-z_j})^{-1} \quad (2)$$

The output of output node m is a weighted sum of its inputs, that is,

$$y_m = \sum_j w_{mj} h_j \quad (3)$$

The input/output mapping function of the ANN may be expressed in vector form as:

$$\underline{y} = \underline{g}(\underline{x}, \underline{\theta}) \quad (4)$$

where: $\underline{x} \in \mathbb{R}^n$ is a vector containing the ANN inputs.
 $\underline{y} \in \mathbb{R}^r$ is a vector containing the ANN outputs.
 $\underline{\theta} \in \mathbb{R}^p$ is a vector containing the ANN weights.
 $\underline{g}(\cdot) : \mathbb{R}^n \times \mathbb{R}^p \rightarrow \mathbb{R}^r$ is a nonlinear vector function.
 n, r, p are the number of ANN inputs, outputs and weights respectively.

It has been formally shown [7], [8] that ANNs with at least one hidden layer (and with a sufficient number of hidden nodes) are able to approximate a wide class of continuous nonlinear functions to within an arbitrarily small error margin. In other words, given a sufficiently large two-layer ANN, it is possible to find a set of weights which will enable the ANN mapping function $\underline{g}(\cdot)$ to mimic virtually any desired mapping function $\underline{f}(\cdot)$. The minimum number of hidden nodes required in order for an ANN to successfully identify a particular function is determined heuristically.

(b) Recursive Training of ANNs

In this application, the ANN is to be implemented on a digital controller. Fig. 2 shows the general block diagram for recursive ANN training. A series of test

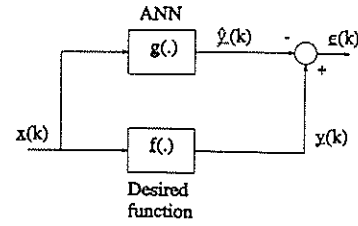


Fig. 2 Recursive Training of ANNs

inputs are presented to both the (possibly unknown) desired function and the ANN. At each timestep k the prediction error $\underline{\varepsilon}(k) = \underline{y}(k) - \underline{\hat{y}}(k)$, may be calculated. The aim of the training algorithm is to use $\underline{\varepsilon}(k)$ to adjust the weights of the ANN so as to minimise the quadratic $Q(k) = \underline{\varepsilon}^T(k) \underline{\varepsilon}(k)$ over time. The *error backpropagation algorithm* [9], used extensively in ANN applications, is a simple algorithm which achieves this minimisation. It is a recursive approximation to a steepest descent algorithm. An equivalent error is calculated for each node in the ANN. The equivalent error δ_m of node m in the output layer is given by:

$$\delta_m(k) = \varepsilon_m(k) = y_m(k) - \hat{y}_m(k) \quad (5)$$

The equivalent error δ_j of node j in the hidden layer is given by:

$$\begin{aligned} \delta_j(k) &= \frac{da(z_j(k))}{dz_j(k)} \sum_m \delta_m(k) w_{mj}(k) \\ &= h_j(k) [1 - h_j(k)] \sum_m \delta_m(k) w_{mj}(k) \end{aligned} \quad (6)$$

Weights connecting the hidden and output layers are adjusted according to:

$$\begin{aligned} \Delta_{mj}(k) &= \gamma_m \Delta_{mj}(k-1) + \gamma_g \delta_m(k) h_j(k) \\ w_{mj}(k+1) &= w_{mj}(k) + \Delta_{mj}(k) \end{aligned} \quad (7)$$

where: γ_m, γ_g are the *momentum* and *gain* parameters respectively.

Weights connecting the inputs to the hidden layer are adjusted according to:

$$\begin{aligned} \Delta_{ji}(k) &= \gamma_m \Delta_{ji}(k-1) + \gamma_g \delta_j(k) x_i(k) \\ w_{ji}(k+1) &= w_{ji}(k) + \Delta_{ji}(k) \end{aligned} \quad (8)$$

The output layer error is calculated first using (5) and then *backpropagated* through the ANN using (6) to calculate the equivalent errors of the hidden nodes. The ANN weights are then adjusted using (7) and (8).

3. NONLINEAR SYSTEM ID USING ANN's

An ideal application for ANNs is in the field of nonlinear system identification [10]. Virtually any discrete-time nonlinear system may be represented by the Nonlinear AutoRegressive Moving Average with eXogenous inputs (NARMAX) model [11], which represents a system in terms of its delayed inputs and outputs. The NARMAX model has the general form:

$$\underline{y}(k+1) = f(\underline{y}(k), \dots, \underline{y}(k-d_y), \underline{u}(k), \dots, \underline{u}(k-d_u)) \quad (9)$$

where: d_u, d_y are the maximum delays in the input and output vectors $\underline{u}, \underline{y}$ respectively.

This form of model is ideal for system identification purposes because the model is expressed entirely in terms of known quantities.

An ANN may be used to identify the NARMAX model of a system by making the arguments of (9) the inputs to the ANN and making the output of the ANN be the one-step-ahead (predicted) output vector $\hat{\underline{y}}(k+1)$ as shown in Fig. 3. TDL denotes a tapped delay line whose outputs are delayed values of its input. The ANN may then be trained to emulate the function $f(\cdot)$ in (9) by comparing the predicted output vector with the actual output vector at time $k+1$, and using the error to update the ANN weights via the error backpropagation algorithm.

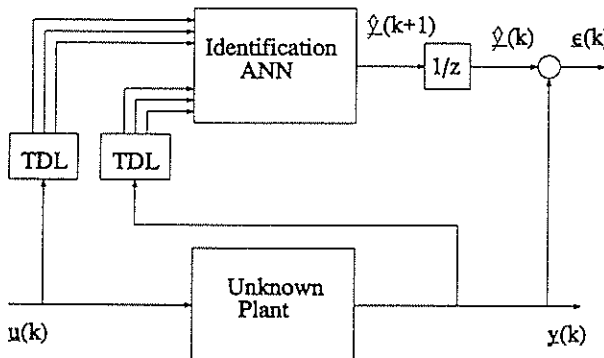


Fig. 3 System identification using ANNs

4. ANN CONTROL OF THE STATOR CURENTS

(a) Electromagnetic Model of the Induction Machine

The discrete-time form of the electromagnetic model of the induction machine, expressed in stationary two-axis (α, β) coordinates is well known and is given by [1]:

$$\underline{i}^{\alpha\beta}(k+1) = \Phi(k) \underline{i}^{\alpha\beta}(k) + c_v \underline{v}^{\alpha\beta}(k) \quad (10)$$

where

$$\Phi(k) = \begin{bmatrix} 1-T/\tau_r & 0 & T\xi/\tau_r & T\xi\omega_r(k) \\ 0 & 1-T/\tau_r & -T\xi\omega_r(k) & T\xi/\tau_r \\ TR_{22} & 0 & 1-T/\tau_r & -T\omega_r(k) \\ 0 & TR_{22} & T\omega_r(k) & 1-T/\tau_r \end{bmatrix}$$

$$\begin{aligned} \underline{i}^{\alpha\beta} &= [i_{\alpha s} \ i_{\beta s} \ \lambda_{\alpha r} \ \lambda_{\beta r}]^T & \tau_r &= L_{22}/R_2 \\ \xi &= L_m/(\sigma L_{11} L_{22}) & \tau_s &= \left[\frac{R_1}{\sigma L_{11}} + \frac{R_2 L_m^2}{\sigma L_{11} L_{22}} \right]^{-1} \\ R_{22} &= L_m R_2 / L_{22} & \sigma &= 1 - \frac{L_m^2}{L_{11} L_{22}} \\ c_v &= T/(\sigma L_{11}) \end{aligned}$$

$L_m, L_{11}, L_{22}, R_1, R_2$ are standard induction machine equivalent circuit parameters.

The dynamics of the electromagnetic system (10) are expressed as four first-order difference equations in terms of the stator currents $i_{\alpha s}, i_{\beta s}$, the rotor flux linkages $\lambda_{\alpha r}, \lambda_{\beta r}$, the stator voltages $v_{\alpha s}, v_{\beta s}$ and the rotor speed ω_r . This model is unsuitable for system identification purposes because the flux linkages are unobservable. It is, however, possible to manipulate these equations into two second-order difference equations expressed entirely in terms of measurable quantities $i_{\alpha s}, i_{\beta s}, v_{\alpha s}, v_{\beta s}, \beta_r, \omega_r$. The derivation of this NARMAX model is beyond the scope of this paper but has the form [11]:

$$\underline{i}_s^{\alpha\beta}(k+1) = f(\underline{i}_s^{\alpha\beta}(k), \underline{i}_s^{\alpha\beta}(k-1), \omega_r(k), \omega_r(k-1), \underline{v}^{\alpha\beta}(k-1)) + c_v \underline{v}^{\alpha\beta}(k) \quad (11)$$

The vector function $f(\cdot): \mathbb{R}^8 \rightarrow \mathbb{R}^2$ is nonlinear with respect to its arguments. Note that $\underline{i}_s^{\alpha\beta}(k+1)$ is a linear function of $\underline{v}^{\alpha\beta}(k)$; this has important consequences on the control method used.

(b) Identification of the Electromagnetic System

The electromagnetic system may be identified by using an ANN to emulate the nonlinear function $f(\cdot)$ in (11). The number of hidden nodes required to identify $f(\cdot)$ was chosen heuristically to be 12. Voltages $\underline{v}^{\alpha\beta}(k)$ are input to the machine and outputs $\underline{i}_s^{\alpha\beta}(k), \omega_r(k)$ are measured. The one-step-ahead predicted current $\hat{\underline{i}}_s^{\alpha\beta}(k+1)$ is delayed by one sample period and then compared with the measured current $\underline{i}_s^{\alpha\beta}(k)$ and the error $\underline{e}(k)$ is used to train the ANN via the error backpropagation algorithm in (5)-(8) as well as to update the estimated value of c_v . The adjustment of c_v is also performed using the backpropagation algorithm, that is, c_v is considered as an additional weight connected to the output layer of the ANN. This training procedure continues in a recursive (on-line) manner. The characteristics of the voltage input

during training is particularly important and in this case Gaussian noise is used to ensure generality.

(c) Control of the Stator Currents

The objective of the stator current control system is to force $\underline{i}_s^{\alpha\beta}$ to follow a desired stator current vector $\underline{i}_s^{\alpha\beta}$. The approach taken in this paper is similar to that used in Model Reference Adaptive Control (MRAC) [2], where the desired output $\underline{i}_s^{\alpha\beta}$ is taken as the output of an asymptotically stable reference model. Control action is then taken to minimise the difference between $\underline{i}_s^{\alpha\beta}$ and $\underline{i}_s^{\alpha\beta}$. This is achieved as follows:

The desired stator current vector $\underline{i}_s^{\alpha\beta}(k+1)$ is calculated at time k . If the nonlinear function $f(\cdot)$ and parameter c_v in (11) are known, then it follows that the input voltage $\underline{v}^{\alpha\beta}(k)$ can be computed from a knowledge of $\underline{i}_s(k)$ and $\omega_r(k)$ as

$$\underline{v}^{\alpha\beta}(k) = [-f(\underline{i}_s^{\alpha\beta}(k), \underline{i}_s^{\alpha\beta}(k-1), \omega_r(k), \omega_r(k-1), \underline{v}^{\alpha\beta}(k-1)) + \underline{i}_s^{\alpha\beta}(k)] / c_v \quad (12)$$

In practice $f(\cdot)$ and c_v are unknown, however, these may be estimated by a system identification ANN as described in section 4(b).

A schematic of the stator current control scheme is shown in Fig. 4. The reference model is chosen as a first-order system with a time constant of 0.01 s, and is expressed in the synchronously rotating (d,q) reference frame. This is to ensure zero controller lag in the stationary reference frame when steady state sinusoidal currents flow in the machine.

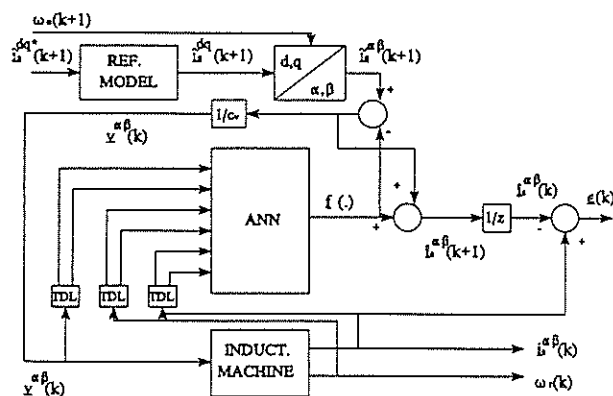


Fig. 4 Adaptive current control using ANNs

The controller of Fig. 4 has two modes of operation; an identification mode, when random voltages are injected into the induction machine and the ANN trained to identify $f(\cdot)$, and a control mode, when the control action is switched in. These two modes of operation are operated sequentially.

(d) Simulation Results

A computer simulation of the current controlled system was constructed. A 2.2 kW induction machine was used. The ANN identifier and controller were simulated as discrete processes with a sampling time of 1 ms. Two of the tests that were conducted on the system are shown. In the first test, the weights of the ANN were initially randomised and the initial estimate of c_v was set to 50% of its true value. Gaussian noise was then injected into the induction machine for two seconds (2000 training steps), and the ANN weights and estimate of c_v adjusted. At a time of 2 s, the control action was switched in, and the response of the system to random steps in i_s^{dq*} and ω_r was investigated. These random steps in the synchronous reference frame correspond to random steps of magnitude, phase and frequency in the stationary reference frame. Fig. 5 shows the actual and desired $\alpha\beta$ currents. After control action is initiated, the actual and desired responses are virtually identical, showing the excellent tracking ability of the system.

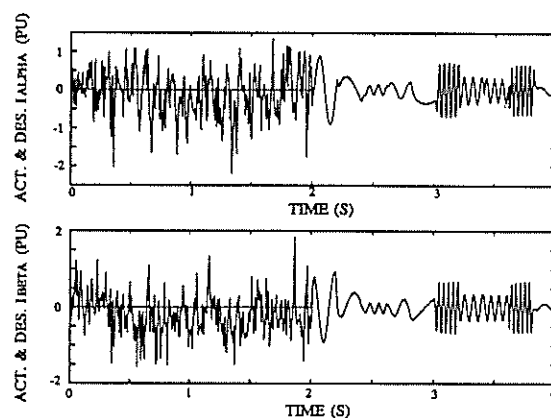


Fig. 5 Training followed by random steps in reference current magnitude, phase and frequency

The second test (in Fig. 6) also shows the response to random steps of magnitude, phase and frequency, but the stator self inductance was halved at $t=0.5$ s in order to investigate the adaptive nature of the controller. The actual responses show small deviations from the desired responses shortly after 0.5 seconds, however these deviations rapidly converge to zero.

These results show that the adaptive current control scheme achieves accurate and stable control and is robust to parameter changes. Furthermore, the controller has been designed with the full nonlinear nature of the induction machine in mind. This system could therefore be used as an autonomous current controller in any induction machine drive regardless of the overall control scheme (eg. vector control, direct self control). It is more robust than many conventional current controllers designed for vector control drives, which are designed assuming perfect vector control in the first place. If the vector control is poor then the performance of the

conventional current controllers deteriorates, whereas the performance of this nonlinear adaptive ANN controller would be unaffected.

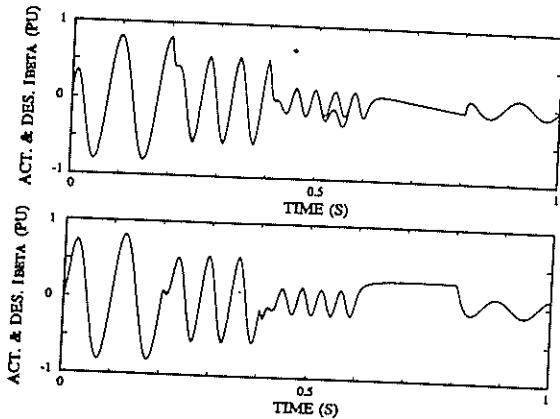


Fig. 6 Response to random steps in reference magnitude, phase and frequency under step parameter variations.

5. ANN ROTOR SPEED CONTROL

(a) The Current-fed Model of the Induction Machine

If the stator currents are controlled sufficiently rapidly and accurately, then the machine may be considered to be fed from a current source. For the design of speed controllers it is convenient to express the current-fed model in the synchronously rotating (d,q) rather than the stationary ($\alpha\beta$) reference frame since all quantities have direct (and not alternating) values. The discrete-time form of the current-fed model, expressed in d,q coordinates, is well-known and is given by:

$$\underline{\lambda}_{r}^{dq}(k+1) = \Phi_c(k) \underline{\lambda}_{r}^{dq}(k) + TR_{22} i_{r}^{dq}(k) \quad (13)$$

where

$$\underline{\lambda}_{r}^{dq} = [\lambda_{dr} \quad \lambda_{qr}]^T$$

$$\Phi_c(k) = \begin{bmatrix} 1 - T/\tau_r & Ts\omega_c(k) \\ -Ts\omega_c(k) & 1 - T/\tau_r \end{bmatrix}$$

$$s\omega_c(k) = \omega_c(k) - \omega_r(k)$$

The mechanical dynamics are described by

$$\omega_r(k+1) = \omega_r(k) + \frac{TL_m}{JL_{22}} (i_{qs}(k)\lambda_{dr}(k) - i_{ds}(k)\lambda_{qr}(k)) - f_{load}(\omega_r(k)) \quad (14)$$

$$= \omega_r(k) + \Delta\omega_r(k)$$

The (possibly unknown) load torque $f_{load}(\cdot)$ is assumed to depend only on $\omega_r(k)$, which is valid for loads such as pumps and fans. For more complex mechanical loads

further delayed speed terms would have to be included.

Equations (13) and (14) describe the dynamics of the current-fed induction machine. In order to convert these equations into equivalent NARMAX form, it is necessary to express (14) entirely in terms of inputs i_{s}^{dq} , $s\omega_c$ and output ω_r . Algebraically, this is difficult (if not impossible) to achieve. It is, nevertheless, known that a NARMAX form does exist. The form of the NARMAX model must therefore be inferred. It is wise, in this case, to estimate the NARMAX model for $\Delta\omega_r$ and not ω_r in (14) in order to incorporate some knowledge of the system into the mechanical equation. Since two electrical states λ_{dr} and λ_{qr} (which are first-order functions of i_{ds} , i_{qs}) are to be eliminated from (14) it is reasonable to assume that this elimination will introduce second-order delay terms, resulting in a NARMAX model of the form

$$\Delta\omega_r(k+1) = f_m(\Delta\omega_r(k), \Delta\omega_r(k-1), i_{s}^{dq}(k), i_{s}^{dq}(k-1), s\omega_c(k), s\omega_c(k-1), \omega_r(k)) \quad (15)$$

(b) Identification of the Current-fed Machine

The current-fed model of the induction machine may be identified by using an ANN to identify the non-linear function $f_m(\cdot): \mathbb{R}^9 \rightarrow \mathbb{R}$ in (15) as discussed in section 3. An ANN with 16 hidden nodes is used to identify $f_m(\cdot)$. Random steps of i_{s}^{dq} and ω_c are input to the current controlled machine and the speed ω_r is measured. The change in speed over one sampling period $\Delta\omega_r$ is calculated and compared with that predicted by the ANN. The error is used to update the ANN weights via the error backpropagation algorithm.

(c) Control of the Rotor Speed

The system in section 4 was linear with respect to its inputs which enabled a direct adaptive control law to be formulated (in (12)). The system in (15) is, however, nonlinear with respect to both its inputs and states, and so no simple direct control law may be found. For nonlinear systems such as this, a system has been proposed [2],[3] which uses two ANNs; one to identify the system, and one to control the system. The ANN identifier (Ni) is first trained to identify the system (as in section B above) and then used as a channel for the backpropagation of errors to the ANN controller (Nc).

Fig. 7 shows a simplified diagram of the speed controller. The controller operates in the synchronously rotating (d,q) reference frame at a sampling rate of 5 ms. The d-axis current reference i_{ds}^* is set equal to the magnetising current. The q-axis current reference i_{qs}^* is set by a conventional speed controller with current limit (the current limit is not shown in Fig. 7). The slip frequency $s\omega_c$ is determined by the adaptive ANN control scheme which operates as follows:

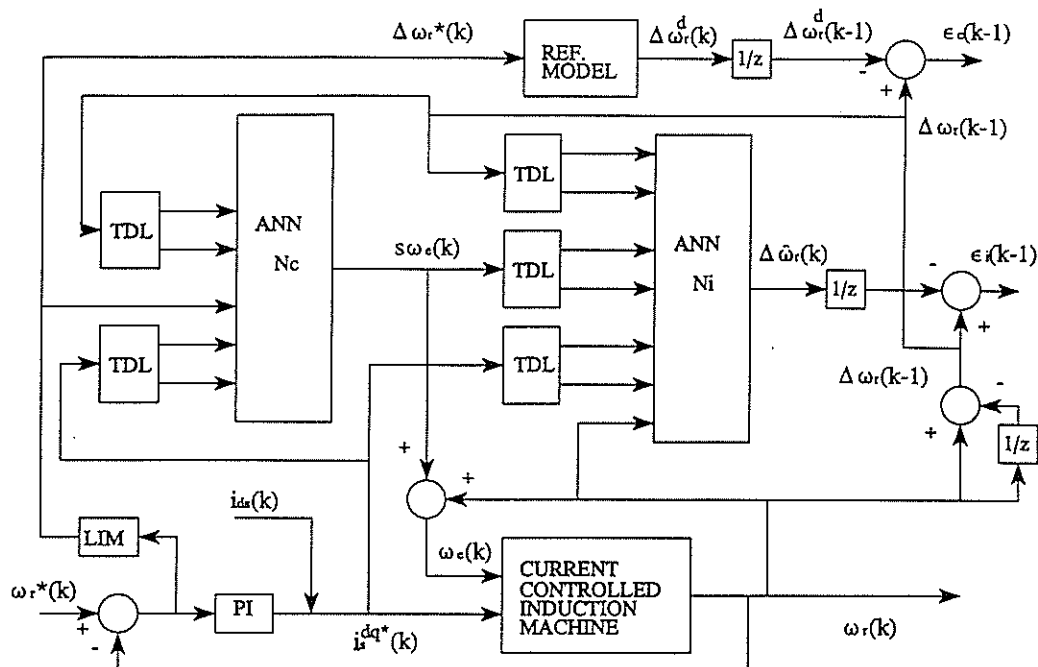


Fig. 7 Adaptive speed control of the induction machine using ANNs

The ANN identifier (Ni) operates as an autonomous system identifier, as described in section B. At each time step a reference $\Delta \omega_r^*$ is calculated as a limited value of the speed error. This reference value is then passed through a reference model to produce a desired change in speed $\Delta \omega_r^d$. This value is compared with the actual $\Delta \omega_r$ to produce a controller error ϵ_c . To train Nc it is necessary to know the error in the slip frequency $s\omega_e$ which caused ϵ_c . This error in slip frequency ϵ_s is obtained by backpropagating ϵ_c through Ni. ANN Nc is then trained by backpropagating ϵ_s through Nc. None of the backpropagation processes are shown in Fig. 7 for simplicity.

(d) Simulation Results

A simulation was constructed of the entire system in Fig. 7, with both adaptive speed and current control schemes. The following sequence of events occurs in order to "commission" the controllers. The current controller is initialised for 2 seconds as described in section 3(d). Random steps in i_{dq}^* and ω_e , changing every 0.5 seconds, are then used to train Ni over a total time period of 20 seconds. Thereafter, random steps in ω_r^* are used to train the slip controller Nc over a time period of 20 seconds. After this initial "commissioning" period of 42 seconds "normal" operation commences. ANNs Ni, Nc and the ANN in the current control system continue training throughout normal operation.

Fig. 8 shows the response to a square wave reference speed ω_r^* of magnitude 200 rad/s (peak-to-peak). Three responses are shown; the speed reference (the square wave), the response of the ANN controlled system (shown as a solid line), and the response of a perfectly tuned vector controlled system with the same current limit (shown as a dashed line).

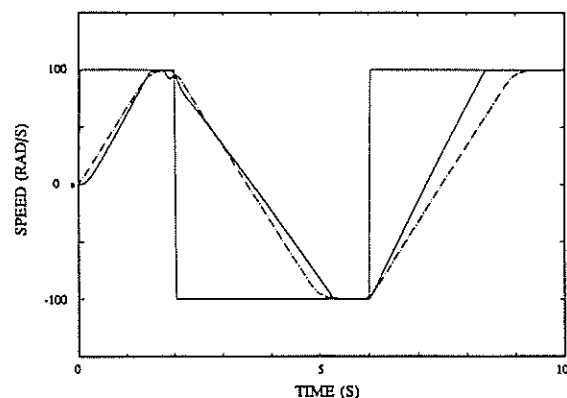


Fig. 8 Response to a square wave reference speed

The response of the ANN controlled system improves with time as the system learns, and during the last transient actually out performs the perfectly tuned vector

control system. The ANN achieves rapid final transient this by running the machine at greater than rated flux. Of course a real machine would saturate at greater than rated flux, however saturation has not yet been modelled in this project. Nevertheless, the responses are stable and suggest that further investigation is worthwhile.

6. CONCLUSION

This paper has proposed the use of artificial neural networks to identify and control the induction machine. Two systems have been considered; a stator current control system, and a rotor speed control system. Both systems are inherently adaptive as well as "self-commissioning". The current controller is a completely general nonlinear controller which can be used together with any drive algorithm.

Both systems show good simulated responses and suggest that further research in this area would be worthwhile. Future work could include:

- a) An investigation into the effect of magnetic saturation on the speed control system.
- b) An investigation of the implications of a practical implementation.
- c) An evaluation of learning algorithms which offer superior performance to error backpropagation.
- d) An investigation into the use of Radial Basis Function (RBF) and recurrent ANNs in place of the feedforward sigmoidal ANNs used in this paper.
- e) An attempt at the development of stability laws for ANN controlled systems.

7. ACKNOWLEDGEMENT

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USING CASCADE NEURAL NETWORKS FOR THE IDENTIFICATION OF RESIDENTIAL LOADS

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Abstract - A previous study by the authors on nonintrusive appliance load monitoring, showed a number of ways in which the features necessary for classifying the appliances, can be extracted. These feature spaces will be summarized in this paper together with suggestions on ways to improve the classification proses. These suggestions include a cascade neural network classification system for reducing the traning set, traning time and hence the classification error.

Introduction

Traditional load research instrumentation has involved intrusive techniques that required the installation of sensors on each of the individual components of the load and the use of complex data-gathering hardware but simple software to determine the usage histories of the appliances.

A more convenient way of load research is by means of nonintrusive techniques where the nature and usage histories of the major appliances in a home may be determined from analysis of the power supplied by the electric utility i.e. aggregate load. The aggregate load is analysed to isolate the characteristic signatures by which the individual appliance can be identified. For example, if the residence contains a refrigerator which consumes 250W and 200VAr then a step increase of that characteristic size indicates that the refrigerator turned on, and a decrease of that size indicates the turn-off event.

Overview of system

Fig 1 shows an overview of the system, of which only the feature extraction phase will be elaborated on. Feature extraction is used to reduce the dimensionality of the space in which classification is done. This also reduces the amount of data needed to train the classifier and thus the training time.

The extraction process does a mapping of the original signal onto a feature space of known and fixed dimension. It is useful to consider a number of different feature extractors and combinations thereof, because the interclass separation properties of the feature spaces may differ considerably.

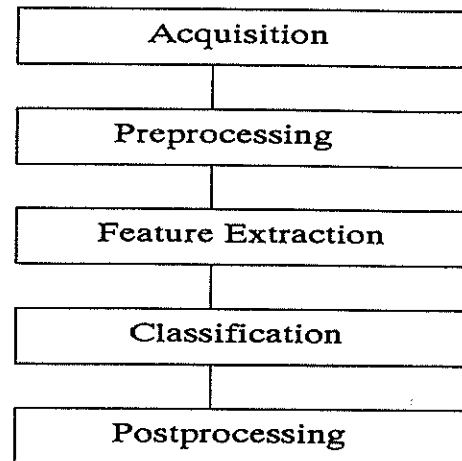


Figure 1: System Overview

Feature spaces considered thus far include:

- Power analysis with the Classical Theory[1]: Based on a *sinusoidal* voltage, a *sinusoidal* current and a *linear static* load condition. Summating the orthogonal power subdivision leads to:

$$S^2 = P^2 + Q^2$$

This maps the entire voltage and current cycle to a single point in $\mathbb{R}^2$ i.e. P and Q;

- Power analysis with Time-domain Correlation[2]: Loading power S is split into active power P and fictitious power F, as in the general time-domain approach. The fictitious power is however further sub-divided into two components namely reactive power and deactive power. The second component of fictitious power, deactive power, is associated with the *non-linearities* and *uncorrelated* waveforms of voltage and current in circuits. It is the result of non-linear characteristics of the load, the source or the network quantities, e.g. impedance of an inductor at different frequencies. Summating the orthogonal components leads to:

$$S^2 = P^2 + F^2 \text{ with } F^2 = Q^2 + D^2$$

This maps the entire voltage and current cycle to a single point in $\mathbb{R}^3$ i.e. P, Q and D;

- Power analysis with the Czarnecki approach[3]: Czarnecki's main contribution to power theory, probably lies in the description of the scattered power component. Active, reactive and scattered power can exist in linear systems. If there are nonlinear elements in the load (i.e. diodes, switches, magnetic saturable materials etc.) then the load current may contain harmonics not present in the source voltage. A fourth power component namely nonharmonic or generated power is thus necessary to fully describe the power contents of any single phase electrical system. Summating the orthogonal components leads to:

$$S^2 = P^2 + Ds^2 + Qr^2 + Dg^2$$

This maps the entire voltage and current cycle to a single point in $\mathbb{R}^4$ i.e. P , Qr , Ds and Dg , with P the real power, Qr the reactive power, Ds the scattered power and Dg the generated power.

- Fourier representation[4]: The current consumed by a load (household appliance) can be represented by a Fourier series expansion. The number of terms represented by the expansion determines the dimension of the feature vector. This method maps the entire voltage and current cycle to a point in $\mathbb{R}^n$, where n is the harmonic order used in the calculations.
- Impedance contour map: Instead of mapping a complete cycle to a single point in a $\mathbb{R}^2$ or $\mathbb{R}^3$ space, a method that preserves the relationship between the voltage and current over the entire cycle must be sought. This relationship is the impedance of the load and is difficult to define for non-linear loads. A description of the contour of the way the impedance changes during a single cycle is a characteristic signature of that load. One way to represent such an impedance curve is by its geometrical moments[5], which have the form of a projection of the impedance image onto the monomial $x^p y^q$.

Decision boundaries with multi-layered perceptrons

In order to understand the motivation behind the idea of a cascade classifier for this problem, the following on multi-layered perceptrons is presented[6].

Backpropagation is in general used as a trainable classifier, where classification in this context means a mapping from a feature space - Euclidean space, usually of high dimension - to the set of class labels, in this case, names of appliances. Trainability means that this mapping is constructed from a set of examples of the mapping desired. To mimic these examples, the classifier creates decision boundaries

which separate the regions assigned to the different classes. The network topology has an important bearing on the decision boundaries that it can create. The larger the feature vector or number of output classes, the more complex and non-linear the decision boundaries. The success of the classification can thus be improved by keeping the feature vector and number of output classes as small as possible. The approximation of the decision boundaries also improves, as more examples are presented in the training phase. Therefore, a smaller feature vector and number of output classes will in general require less training samples for the same decision boundary approximation.

Cascade networks

A very large range or diversity of household appliances exists. Not only do we encounter for a specific type of appliance a large range of different types eg. infrared heaters with glass elements, oil heaters, resistance heaters with fan, etc., but different operating conditions (states) and combinations thereof for a specific appliance exists. For example, a hair dryer can have fast or slow and warm or cold settings. Each combination leading to different power consumption and thus a different characteristic signature by which it can be identified.

The feature vector to discriminate between these appliances, and the network architecture for classifying it, can become very complex if all the different types of appliances, each with different operating state combinations, is modeled as separate classes of a single system. This also requires a large training set, which again results in long training times.

One way to reduce the vast range of classes, is for example to group all the large and small hair dryers, as well as their fast, slow, warm and cold settings into a single hair dryer class. It will then depend upon the network classifier to construct the complex non-linear decision boundaries necessary to distinguish between the different classes.

Another, and more intuitive way to reduce the number of classes in the classification process, is to use the natural grouping of household appliances into a few general types eg.:

- Resistive appliances: They form the largest category, encompassing heating appliances such as kettles, ovens and heating elements of washing machines and dishwashers. These appliances are characterized by zero reactive power and the absence of harmonic current components. Fig.2 shows the impedance curve over a single voltage cycle of a kettle.
- Pump/Compressor-operated appliances: These are electric motors driving a pump. Several electrical household appliances enter this category, in particular: refrigerators, freezers, and dishwasher and washing machine drain pumps. These appliances are generally characterized by

substantial reactive power and odd-numbered harmonic current. The impedance curve over a single cycle of a refrigerator is shown in Fig. 3.

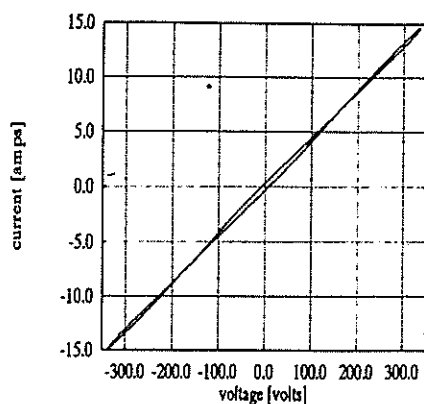


Figure 2: Kettle - resistive

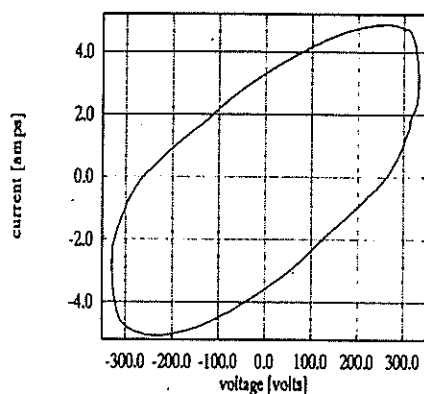


Figure 3: Refrigerator - pump-operated

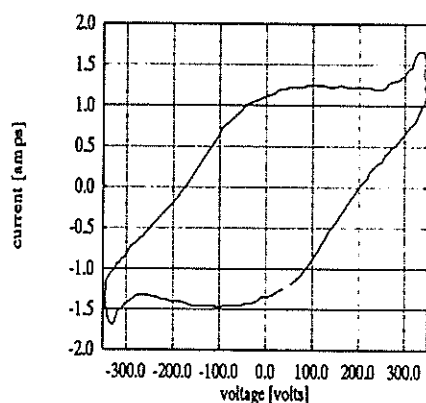


Figure 4: Fan - motor-driven

- Motor-driven appliances: This category encompasses the other appliances bearing an electric motor such as a washing-machine, fans, hair dryers and mixers. These appliances differ

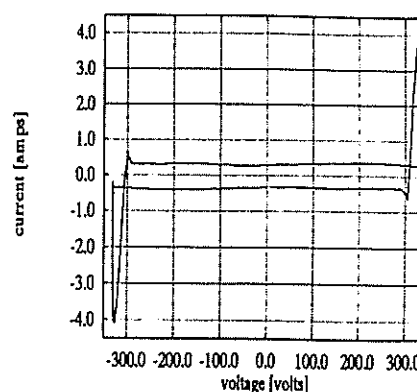


Figure 5: Colour Television - electronically-fed

from pump-operated appliances by their generally less substantial switching-on transients. Fig 4 shows the impedance curve of a fan, taken over a single voltage cycle.

- Electronically-fed appliances: This category comprises low consumption appliances such as televisions and personal computers. They are characterized by a current spectrum rich in harmonic components. Fig. 5 shows the impedance curve of a television.
- Fluorescent lighting: This type of lighting is characterized by very high generation of the third harmonic of the current and substantial current-voltage phase shift.

The first stage of the classification will therefore be the identification of the appliance by type, while the second stage will identify the type of appliance by name. An advantage of such a separation is that the different sub-networks of the two stages can be trained individually on a feature set which is optimal for that specific network. For example, the first stage network will use only the minimum features necessary for classifying the appliance by type, while a network in the second stage, for instance the network for naming motor-driven appliances, will only use features to discriminate between such appliances, and not all appliances. Each network will then have a feature set of much reduced dimension, fewer output classes and network architecture of reduced complexity. This in turn will give rise to decision boundaries that can be approximated more easily and successfully, and with fewer training samples and time.

Results

The training set consisted of 520 feature vectors provided by 28 different appliances from 2 different houses. The test set consisted of 70 different feature vectors selected randomly. The first results were obtained through the use of a multilayer perceptron classifier with one hidden layer. Backpropagation

with conjugate-gradient optimization was used for its training. The following table shows the best classification on the test set obtained by using a 3 layer neural network. Only a single network was used for classification.

Feature space	Input neurons	Hidden neurons	%
Classical	4	20	68.12
Correlation	5	32	77.2
Czarnecki	6	40	82.0
Fourier	7	25	88.31
Impedance moments	12	40	88.41

These results are by no means representative of the overall system and are only presented to evaluate the different feature spaces. For example, instead of using one class for all hair dryers, it can be subdivided into different classes e.g. small and large hair dryers. This way of labelling will most certainly have an effect on the classification.

When a cascade network topology was used for classification, an increase in the success rate was found in general, apart from the decrease in training set size and time, which shows that the classification can be improved by sub-dividing the classification system into several sub-systems or cascade classifiers.

Conclusion

At the present stage of the study, the belief is held that household loads can be monitored in a nonintrusive way. The study presented in this paper shows how a cascade network topology can be used to first classify the unknown appliance by type (a natural grouping found among all types of domestic appliances) and the second network classifies the appliance by name.

Future work will involve research into feature spaces that are more specific, and thus optimal, for the different types of appliances (sub-networks), in contrast to the present, more general, feature spaces. Preliminary results show that a cascade configuration can lead to an improvement in classification, due to the simpler network architectures, decision boundaries and reduced feature vector and number of output classes.

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THE COMMISSIONING AND PERFORMANCE OF A NEW INSULATOR POLLUTION MONITORING APPARATUS

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1. ABSTRACT:

A new insulator pollution monitoring apparatus has been developed at the University of Stellenbosch. This paper describes the construction and performance of the device. The physical and electrical characteristics of the device are specified and explained. The philosophy used to overcome transformer inrush magnetization is explained. Furthermore, the paper describes the calibration procedure and the results obtained, showing how the results obtained can be interpreted. The paper also discusses the practical applications for IPMA and the direction of further research.

2. INTRODUCTION:

The University of Stellenbosch has previously been involved in the measurement of the pollution severity on high voltage insulators. In this regard, a semi-portable, self contained insulator test station was designed and commissioned in 1989/1990.[1] The work detailed in this document is the continuation of this line of research.

The IPMA has been designed as self contained test station, based on the principle of surface conductance measurement.[2] It encompasses test insulators, HV transformer, steam generator, lid raise/lower motor and a control and data logging module. The entire device is controlled by a computer program, and requires only a 220V, 50Hz supply and a water point. A weather monitoring station is also needed for meaningful interpretation of the results.

The monitoring apparatus has undergone changes: both physical and operational. These, it is hoped will give more coherent data than was achieved from the previous Insulator Pollution Monitoring Apparatus (IPMA). To this end, the control of the operation and the interpretation of the results have been overhauled. The performance of the IPMA has been evaluated in the laboratory and the calibration data obtained will be used to evaluate the performance of the device in the

field.

3. CONSTRUCTION:

The IPMA consists of a polymer cabinet mounted on a galvanised steel frame. The cabinet contains a 220/1100 V single phase transformer, a steam generator, the shell motor and various relay switches used for controlling the device. Above the cabinet is a T-bar construction allowing for the suspension of two separate insulators or strings of insulators. A moveable shell, driven by a helical screw can be moved up and down to cover either the insulators, for the duration of a test, or the base cabinet between tests.

The most obvious difference between the new version of IPMA and its predecessor is the reduction in physical size. The new IPMA is designed to be mounted on the standard ESKOM half-cradle, whereas the previous unit was a much larger floor mounted affair. One of the main reasons for this is the reduction in operating voltage from 11 kV to 1.1 kV. This has obviously led to the use of a smaller transformer, and clearances around the high voltage terminals have also been greatly reduced.

Another significant reduction in size was achieved by the use of a pressure steam generator operating at 1.5 bar as opposed to the steam kettle used in the previous device. This has led to a difference in the mode of operation, but this will be detailed below.

4. PRINCIPLE OF OPERATION:

The IPMA operates by measuring the conductance of the pollution layer of a given insulator in order to make a prediction of the degree of pollution severity present.[2]

The measurement is performed by raising the shell, generating steam and measuring the peak current that flows when the insulator is energized for a period of a few cycles. More steam is generated, the transformer is energized, and the next value of peak

leakage current is measured. The output variables from the IPMA, two currents and the supply voltage, are logged onto the hard disk of the controlling computer for later retrieval and analysis.

5. OPERATION ROUTINE:

The exact operation of the IPMA is governed by the following operation routine:

1. Raise moveable shell to enclose insulators.
2. Dry insulators to ensure equivalent initial conditions for all tests.
3. Generate steam and wait for *Steam Ready* signal.
4. Energise transformer and log maximum leakage currents on both insulators.
5. Release steam.
6. Energise transformer and log maximum currents on both insulators.
7. Wait for *Steam Ready* signal.
8. Repeat steps 5. - 7. until maximum current measured is less than 95 percent of the highest current measured for the duration of the test.
9. Dry insulators by means of hot air blowers.
10. Lower shell to expose insulators to ambient conditions between tests.

The following parameters may be set in the controlling computer setup routine which affect the operation of the IPMA:

The transformer may be left in an energized state during the period between tests to investigate the effect of the formation of dry bands on pollution accumulation.

The condition to end the test may be set such that the test sequence is ended when either of the insulator currents begins to decrease, or when the current on a particular insulator decreases.

Furthermore, the length of time that the hot air blowers are on and that the steam valve is open is specified in the controlling computer setup routine. The number of tests per day, and the times at which these are to occur are also specified in this way. The duration of the transformer energization is controlled by hardware, and has to be physically adjusted on the instrument.

Date	Time	
Vsupply	I ₁	I ₂
-	-	-
-	-	-
-	-	-

Figure 1: Format of IPMA test output file.

The controlling computer generates files containing the results of the tests in the following format (see Figure 1).

The data is stored in ASCII format and provision has been made for sufficient data storage to allow for intervals of up to six months between site visits for the purpose of data retrieval, assuming that two tests per day are performed.

6. CALIBRATION RESULTS:

The IPMA has been calibrated in the laboratory, prior to installation on site. The rationale for this laboratory calibration is to verify the possibility of the concept, and to enable a comparison to be drawn between artificial and real results obtained by IPMA. The artificial tests were performed by polluting standard U70 cap and pin insulators with a solution of kaolin and salt water mixed in accordance with IEC guidelines.[3]

The calibration procedure consisted of two parts. The first involved examining the waveforms of current and voltage obtained during the test to ensure that these were applicable to the method envisaged for the field device. Once this had been established, the captured data was used to determine the values of conductance associated with the insulator being tested. Thereafter, equivalent salt deposit density (ESDD) measurements were performed and the results plotted.

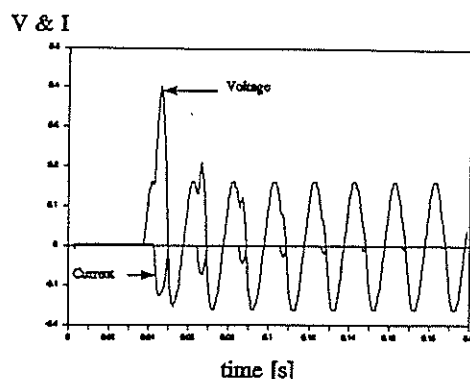


Figure 2 Transformer voltage and current showing effect of magnetizing inrush

Initially problems were experienced due to non-sinusoidal voltage and current waveforms as shown in figure 2.

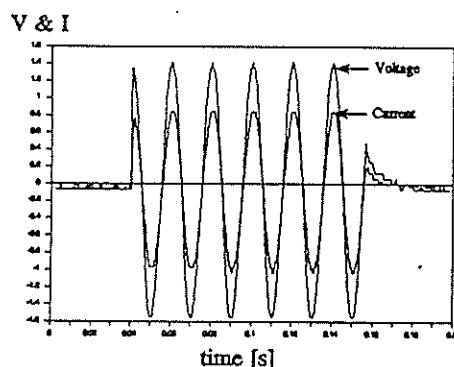


Figure 3 Transformer voltage and current with switching at voltage maximum.

This problem can be ascribed to the large transformer magnetizing inrush current which occurs if energization occurs at a zero crossing of voltage. A thyristor circuit was therefore developed in order to perform switching at voltage maximum. The waveforms captured after this modification can be seen in figure 3.

The peak values of the current and voltage obtained

from the test waveforms are used to calculate the conductance of the pollution layer on the insulator. As a first approximation, a linear regression was applied to the data to yield the curve shown in figure 4.

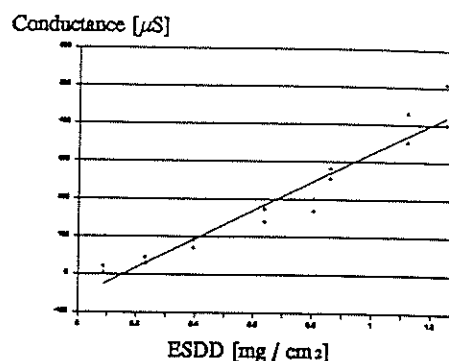


Figure 4 Linear regression on IPMA calibration data.

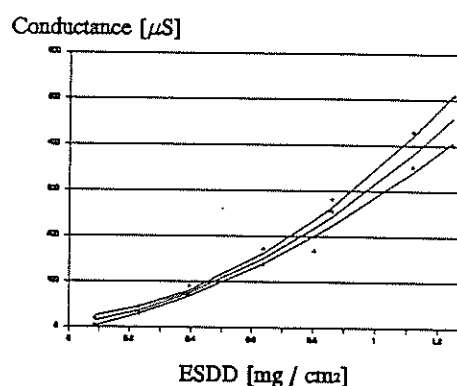


Figure 5 Parabolic regression on IPMA calibration data.

A better representation of the test data is obtained by applying a parabolic function to the test data as shown in figure 5. The centre line indicates the best fit parabola for all the data points. The upper and lower lines show the best fit parabolas for the upper and lower data points at each value of ESDD respectively.

7. INTERPRETATION OF RESULTS:

Based on the curves depicted in figure 4 and figure 5, it would appear that the relationship between conductance and ESDD can be reasonably approximated by a parabolic curve, particularly where the ESDD falls into the category of *moderately severe* or higher. The approximation of the relationship as a straight line is not as accurate, but it is probably adequate as a first approximation. A possible reason for the non-linearity observed is that the ESDD is not

a linear function of the conductivity of the solution.

The coefficients of the equations of the various curves are given in the following table:

Table 1 Coefficients of best fit curves for IPMA calibration results.

Function	a	b	c
Linear	0	355.54	-25.7
Best Fit Parabola	218.85	91.18	4.82
Low Limit Parabola	162.61	131.53	-9.10
High Limit Parabola	275.10	50.82	18.74

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8. FUTURE IMPLEMENTATION:

On completion of the laboratory calibration procedure, the IPMA is to be installed at Koeberg Insulator Pollution Test Station (KIPTS). The data obtained from IPMA will then be compared to that obtained from continuous leakage current measurements, half-hour peak insulator leakage current measurements, weather data and the contents of dust deposit gauges. It will also be possible to periodically compare the IPMA data with ESDD measurements obtained from insulators installed at KIPTS.

The performance of the test insulators can also be made generally applicable by converting the conductance values measured to conductivity values by means of the form factor of the insulator. In this way, various profiles of ceramic as well as polymeric insulators can be evaluated. It will also be possible to compare the results of insulators tested by the IPMA with those obtained from other standard tests such as the clean fog test.[4]

Once a number of these IPMA devices have been strategically located around South Africa, and the data has been evaluated, it will be possible to generate a pollution severity model for the entire region.

9. REFERENCES:

AN ANALYTICAL MODEL OF SURFACE ELECTRICAL BREAKDOWN IN SOLID INSULATION MATERIALS FOR POWER ELECTRICAL EQUIPMENT IN HUMID AND CHEMICALLY POLLUTED ATMOSPHERES

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It is a well-known fact, that the presence of aggressive gaseous pollutions of the atmosphere of industrial sites leads to accelerated deterioration and the complete destruction of electrical equipment. The surrounding environment of chemical plants, some installations of the machine-building, the oil-refining, the food and the agricultural industries contain significant quantities of aggressive gases, dust and ashes. The adverse deposits often reach for some townships tens of t/km².month. Estimations for Duisburg, Liverpool and Pittsburg are about 50t/km².month. The air pollution over London in terms of sulphuric acid is about 1 000 - 2 000 tons. [1]

The effect of air contaminants especially on electric insulation materials in current use usually causes a drastic decrease in their resistance to breakdown by surface discharges. The surface electrical breakdown (tracking) strength specifies to a considerable degree in particular their applicability for heavy duty under voltage in adverse environments and is essential for dimensioning the surface insulation gap spacing. Forced-over dimensioning of insulation constructions caused by the deterioration of the polymers' surface electrical breakdown strength (SEBS) and the imposing miniaturisation trend in electrical equipment design require a thorough analysis of the problem all the more that the number of electrical high-voltage equipment operating in adverse industrial contaminated areas exceed 25-30% of the total number of the utilised technical devices. The maximum time-period of operation for some electrical equipment in chemical production sites before capital repair does not exceed 12 to 17 months. Numerous investigations in natural conditions emphasise the joint effect of climatic factors and locally concentrated gaseous components.

1. Theoretical background of the surface breakdown in polymers exposed to humid and contaminated atmospheres.

In the presence of gases and steam in the environment of a polymer, mainly pure diffusion processes are usually considered to form the initial phase of their interaction. The established diffusion layer of dissociated external particles inside the material possesses strongly increased ionic conductivity γ .

The solution of the known diffusion wave equation presents the concentration c of the

external medium in the diffusion zone of the material, defined by two points of motion of the wave : ($x = 0; t_D = 0$) - on the surface and (x, t_D) inside the material:

$$c = c_2 \left(1 - \frac{2}{\sqrt{\pi}} \int_{z=0}^z e^{-z^2} . dz \right) \quad (1)$$

where t_D - time period of diffusion,

c_2 - saturation concentration in the material,

$z = \frac{x}{\sqrt{Dt_D}}$ - wave parameter

D - diffusion coefficient of the material.

Based on the concept of a thermal - type breakdown mechanism, Wagner's et al. stability criterion of the already established thermal channel after applying an electric field for the time t is being used [2]:

$$E_{bd} \cdot \gamma_{in} = \frac{4Q}{\beta \cdot r_{out}} \quad (2)$$

where γ_{in} - initial conductivity of the material (at initial temperature T_0),

Q - chemical conductivity of the material,

r_{out} - outer radius of the heat channel at end stationary state and

β - material constant in the range $T - T_0$.

Assuming 'stationarity' of the diffusion wave i.e. of point 'x' relative to the much faster thermal processes it might be possible to introduce into the expression of the stability criterion the conductivity γ . The following expression for the outer radius of the established heat channel, which enters the material's diffusion zone is accepted and hence, for the breakdown diffusion time function:

$$r_o = \frac{x}{2} = \sqrt{D} \cdot z \cdot \sqrt{t_D} \quad (3)$$

The critical wave parameter Z_{cr} (to realise the first possible breakdown is time-independent: $Z_{cr} = 1.225$. From (2), for $\varphi(z) = 0,917$ and $C_{cr} = 0.083$:

$$E_{bd}^2 \cdot t_D \approx const \quad (4)$$

which gives the thermal breakdown in the subsurface layers of the material, below the breakdown value of air, due to the penetration of external ions by way of diffusion.

The equations of the mentioned type consider only the diffusion in solid dielectrics. The present experimental data clearly show a significant non-compliance with (4).

To investigate the mechanism of the electrical deterioration of solid insulating materials used in high voltage equipment, the electrical behaviour of 15 different polymer materials

(laminates and composites) has been studied in chemically aggressive and humid atmosphere. The change in the surface breakdown voltage to the relevant exposure time-period was chosen to characterise the dielectric surface failure strength of the materials.

2. Methods used in the study of the materials exposed to adverse humid atmospheres.

The extremely thorough analysis of possibilities for simulating the adverse effect of contamination and increased air humidity on high-voltage insulating materials has finalised in the important conclusion that to produce a degradation effect in the surface layers of the material only two test media (forming on principle a strong and a weak electrolyte) should be needed. The more moderate gaseous concentrations are used for qualitative comparisons and possess a practical significance - they are close to the admissible sanitary norms. The results are compared with those in a pure humid atmosphere.

After a specified exposure time in the chamber, the samples (with normed dimensions) have been tested with the applied, smoothly risen, 50Hz, sinusoidal voltage until flashover, or breakdown failure. The choice of test voltage and the type of electrodes used is motivated by the aim of the investigation and their combinations are most diverse [1]. The electrodes used are chosen to cover most of the cases in equipment design. To establish the parameters of the potential field between the electrodes as well as to show the effect of their configuration on the distribution of the equipotential lines, the method of modelling of potential fields on conductive paper was applied. The drawing-up of the field (with axial symmetry) was made by means of the conform images method.

To analyse the effect of the adverse atmospheres on the structure of the subsurface layers of the polymers and to explain the degradation processes which motivate the fall down in their surface breakdown strength, the infrared spectra of the pure resins and their respective end-products have been recorded.

To follow the relatively fast mechanism of arc firing and its progress between the electrodes, breakdowns on clean and dry samples and on such after exposure in pure humid and in 70 mg/l concentrated nitrogen oxides humid atmospheres have been recorded by means of a super fast camera (20 000 shots/s)

3. Analysis of the experimental results

3.1 Behaviour in pure humid atmospheres

Data from infrared spectra show no effect on the resins - apparently there are no signs of hydrolysis with the exception of the cotton fabric bond laminate (with phenolcresolformaldehyd resin), very intense at the end of the 30 days test

period. The intensity of OH -groups increases about the 4 - 5th day. The same effect, but less intense is being observed with the composite material on melaminephenolformaldehyd resin basis. The decrease in breakdown voltage V_f may find an explanation in the absorbed moisture (the 'spilling' of the characteristic peak at $3\ 400\text{ cm}^{-1}$ - $3\ 500\text{ cm}^{-1}$). The same rise in the bands at $1\ 100\text{ cm}^{-1}$ may be observed with the bakelite-type materials: Breakdown (failure) -between the 9th and the 14th day of exposure.

3.2 Behaviour in humid atmospheres with SO_2

Melamine - based materials show no change at SO_2 concentrations of 10 mg/l as well as no significant changes in V_f (flash-over) curves. Some SO_4 groups are appearing at 100 mg/l. No visible destruction of the melamine ring can be observed and the changes in general are small. Barely about the 30th day a slight sulphonization of the ring occurs and the flash-over voltage decreases.

With the glass-fabric epoxy bond laminate and the epoxy resin mouldings no changes in the spectrum can be seen. The slight fall-down in the curves about the 30th day should be explained by the existence OH -groups in the maximum at 3340 cm^{-1} .

The bakelite-type materials react intensively on both SO_2 - concentrations with the presence of H_2SO_3 and SO_4 groups, related probably to a mechanical connection between the fillers and the appearing sulphurous and sulphuric acids (strong electrolytes).

The best examples are the laminated paper-bond materials with quick sulphonizing of cresol and increase in OH -groups. Full fall-out in 20 days (fig. 1).

3.3 Behaviour in humid atmospheres with nitric oxides.

With the exception of the melamineformaldehyd based composites and partially-epoxy mouldings, there is no material without failure. Decreases in flash-over voltages are considerable (for cotton-fabric and paper-bond laminates - up to 80%; bakelite-phenolformaldehyd-resin - 50 to 75%), which is to be related to a certain nitration of cresol and phenol, disruption of the CH_2-O-CH_2 bridges and the connection of NO_2 - groups to the acyclic chain. (fig. 2) Breakdown in bakelites may be explained by their strong hydrophilic properties leading to intense sample moistening and penetration of nitic electrolyte. With the glass-fibre laminates the process is aggravated by diffusion into the voids between the epoxy resin and the glass fabric.

3.4 Arc topography

The values of the electric permittivities, ϵ_r , of the tested materials vary from 4,43 up to 24,46 and are relatively low in value. The surface flashover voltages are high (about 8 kV at 10 mm interelectrode distance). The field is relatively undistorted by the difference in permittivities of air and the material. The 'border' effect of capacitive currents is negligible. The surface layers of the polymers possess good or acceptable dielectric properties. The flashover (material dependent) is realised at different proximities to the sample's surface but always in air. Later on, due to electrodynamic forces, the arc is blown-up in the characteristic way (slower move of the anode, passage through zero, restriking) to steady burning on top of the electrodes. Nevertheless, the solid introduces a certain non-uniformity into the field and the flashover is initiated at lower values. With exposure to pure humid atmospheres and subsequent drying-out the development observed is similar. After 20-30 days of exposure to nitric oxides and high air humidity, ionic products of interaction are formed in its subsurface layers initiating migration polarization conditions with a rise in ϵ -values 3 to 14 times. The values of the initial flashover decrease significantly. The conductivity in those layers is high and the active currents there also increase. The arc ignites immediately by or on the surface whereupon it follows the described way (fig. 3). A maximum approximation based on a computer program with 18 mathematical expressions (linear, quadratic, polynomial, etc) has been found for the experimental curves $V_f(t)$ of a composite (resolphenolanilinaldehyde resin) with intermediate properties. The function $y = BA^x$ gives the lowest standard deviation. To assess the effect of the experimental factors on the breakdown resistance, the method of the dispersion multifactor analysis has been applied for 3 materials: Withstanding, with average characteristics and non-withstanding.

4. Analytical discussion of the experimental data based on the approximation curve $y = BA^x$ and its comparison with the quadratic expression $E_{bd}^2 \cdot t_D = const.$

The analysis of the changes in the tested materials and their electrical behaviour in adverse media leads to the following conclusions:

- a) Stages of influence by adverse gaseous media
 - the formation of a continuous electrolyte layer on the surface of the polymer.
 - diffusion of the electrolyte into the subsurface parts
 - chemical interaction between the aggressive

medium and the material to irreversible changes in the structure of the latter.

- b) Increase in the polymers' conductance, caused by:
 - increased number of ions of external origin (mostly $\leftrightarrow OH$ -groups) on and in the material
 - destruction of the basic molecules with the formation of acidic groups chains disruption etc.
- c) The two concentrations of the aggressive gases used lead in fact to the same characteristic changes having but different intensities.

As it was already mentioned, the modelling breakdown function $E_{bd}^2 \cdot t_D = const$ seems reasonable only in the presence of pure diffusion and allows only qualitative considerations. The practical quantitative assessment of $c_1 = c_2[1 - \phi(z)]$ due to its dependence on many factors is very difficult and no numeric values for $E_{bd}^2 \cdot t_D = const$ are yet known. Moreover, in the experimental practice, the time t' of exposure is usually used, for which the function of breakdown voltage change is unknown. (compared to the critical time of the diffusion wave penetration t_D). In fact both times are not identical).

The function $V_f^2 \cdot t'$ for three representative materials after exposure to steam (classic diffusion) is shown in fig. 4. Analysing the respective spectrograms it seems that:

- the function increases, apparently there are not any hydrolytic processes, the decrease in V_f is due most probably to absorbed surface water, but not to any inward diffusion: $V_f^2 \cdot t' \neq const$

- from the 6th day on (with the bad material-already after 3-6 days) a very

distinct band rise at 1600 cm^{-1} (characteristic for absorbed water) can be seen. The first failures show up too:

$V_f^2 \cdot t' \approx const$. The diffusion is relevant, i.e. the exposure time t' is adequate for reaching the critical value of t_D . Although slowly, irreversible changes are initiated.

- the decrease in $V_f^2 \cdot t'$ at the end of the exposure period and the abrupt intensification in failures is obviously due to hydrolysis (pronounced rise in the number of OH^- groups at 3420 cm^{-1})

With the other gas media (especially nitrogen) the principle of the three phases remains valid, taking of course into consideration the respective gas dissociation constant. The degradation of the materials is accomplished sooner or is more intense due to their destruction and the formation

Fig. 4

Duration of exposure (days)	Type of aggressive agent			Material (resin)
	SO ₂	N ₂ O ₅	H ₂ O	
1	105.5	78.4	120	Resol-phenol-aniline-formaldehyde
3	279.0	104.0	308	
6	483.0	318.0	591	
9	723.0	426.0	690	
14	992.0	520.0	805	
20	1210.0	500.0	840	
30	154.0	480.0	750	Phenol-formaldehyde
1	102.0	50.2	110	
3	248.0	72.0	225	
6	252.0	46.9	266	Melamine-phenol-formaldehyde
9	143.0	-	144	
1	95.5	81.0	133	
3	292.0	147.0	384	
6	545.0	187.0	750	
9	790.0	182.0	1065	
14	1260.0	134.0	1485	
20	1867.0	-	1880	
30	2330.0	-	2430	

of additional ionic groups. The total duration of the 3 phase is shorter. In the case of nitric oxides both times t_D and t' are even in the same range. (laminates).

CONCLUSIONS

According to the analysis of the experimental results gained by means of an accelerated testing, the cause of the decrease in the level of the surface dielectric resistance in terms of flashover voltages as well as in the appearance of breakdown failures is caused by :

- the existence of an "external" electrolytic layer, physically connected (attached) to the material surface, and
- an internal, subsurface conductive layer. The latter is determined by the equilibrium condition of the diffusion into the material's subsurface layers and the degree in irreversible changes in the structure of that layer.

The changes in the breakdown voltages do not follow the relation $E_{bd}^2 \cdot t_D = \text{const}$. It seems possible for practical reasons to assess a final degradation level of the widely used in high-voltage equipment insulation materials by means of an accelerated test method and to classify them for use in adverse atmospheres. Based on the wide range of experimental results and applying the laws of chemical kinetics life

expectancy has been mathematically assessed an expression for the relations for the changes in a particular characteristic (life duration) s to a pre-given value s' in relation to the temperature and the concentration of the medium $\sum c_{aj}$ and $\sum c_{aj}$ has been derived.

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- [1] Gochev B., On the methodology of modelling the effect of an aggressive medium on the surface dielectric strength of solid insulating materials, II International conference ELISOT, Varna 1982.
- [2] Ganev L., Gochev B, Sarafov S. A method for prognostication of the behaviour of polymer materials in adverse gaseous media, So.Soss of IEP. Vol. 4, Sofia 1985.

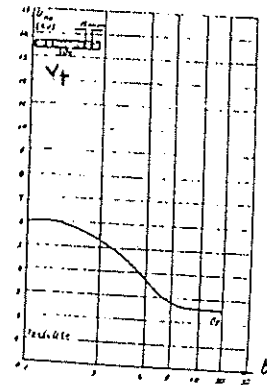


Fig. 1

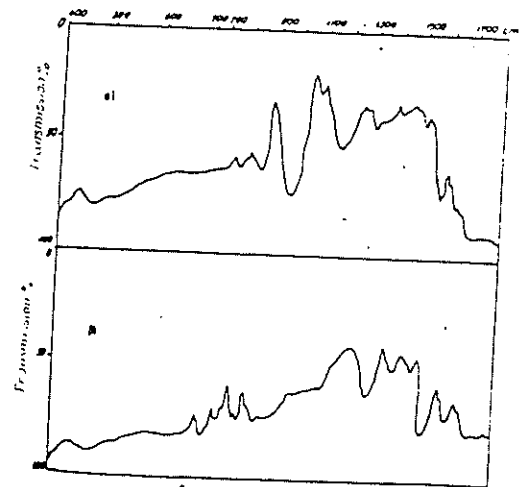


Fig. 2

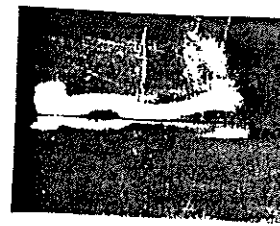


fig.3a

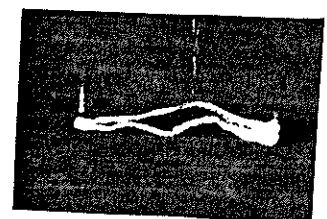


fig.3b

DISCRIMINATION BETWEEN ELECTROMAGNETIC NOISE GENERATED BY SUGAR CANE FIRES, CONDUCTOR CORONA AND POLLUTED INSULATORS ON OVERHEAD LINES

A C Britten[†]J P Holtzhausen[†]W L Vosloo[‡]C R Evert[†]

Abstract : The paper briefly explains the need for the detection of sugar cane fires under 275 and 400kV lines in Natal in particular. The excitation function concept is applied to the translation of cage measurements of fire and conductor corona noise (all quasi-peak data) to an elementary single phase line geometry. Together with summated radio noise currents from polluted insulators, discrimination between the three noise sources is analysed as a function of line length and weather conditions.

1. INTRODUCTION

Recent research done by the University of Stellenbosch and Eskom has shown, on the basis of corona cage tests, that it is possible to discriminate between the magnitudes of electromagnetic noise generated by conductor corona (for both wet and dry conditions) and simulated sugar cane fires (1, 2, 6).

The electromagnetic noise here refers to both band - limited quasi-peak levels and broad band time domain measurements of noise pulse amplitudes, repetition rates and position relative to the applied power frequency voltage.

The paper sets out to describe a simple procedure by which cage noise levels (measured on a quasi-peak basis at a centre frequency of 500 kHz) may be transformed to the levels occurring on an elementary single phase transmission line. From the above it is shown how the propagation of conductor corona and fire noise may be modelled, and how this makes it possible to discriminate between the two sources. Transformations to three phase lines are under consideration.

It is also shown how the above concepts are applied to the modelling of the propagation of radio noise generated by polluted insulators. This makes it possible to explore the discrimination between all three sources of noise.

2. SOME PRACTICAL BACKGROUND

For several years, widespread burning of sugar cane in Natal caused extraneous phase-to-phase and phase-to-ground faults in a large network of 275 and 400kV lines. These faults have been, and still are, very disruptive to the network itself and various large industrial loads.

One of the strategies being developed to avoid or to mitigate the effects of flashovers is the reliable detection of a fire under a line prior to flashover. This is intended

to make it possible for the line to be timeously de-energised or for its auto-reclose protection to be temporarily disabled.

The Cage research and some field measurements already done have shown that the presence of a fire near energised conductors causes high levels of corona-generated radio noise (1, 2). Practical research is now being done on the detection of such noise in operational power line carrier circuits : hence the need for the discrimination study.

3. THEORETICAL CONSIDERATIONS

Consider the elementary single wire transmission line shown in figure 1 below.

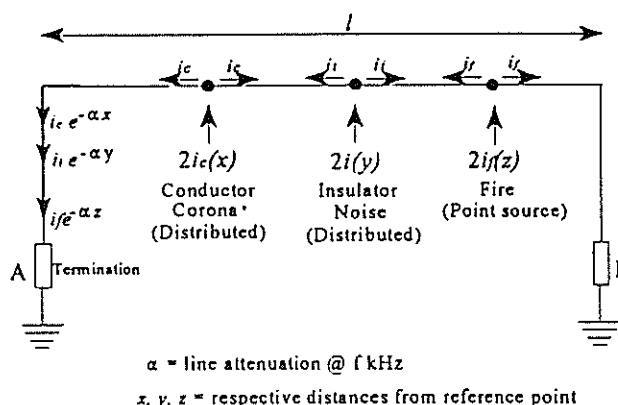


Figure 1 Elementary line showing various noise sources.

3.1 Conductor Corona Noise

It is assumed that infinitely many sources of corona current are distributed along the conductor shown in figure 1 and that their amplitudes are not correlated in time. The total propagated current I arriving at termination A is therefore found by quadratic summation.

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$$I = \sqrt{\int_0^l (i_c e^{-\alpha x})^2 dx} \quad (1)$$

$$= \frac{i_c}{\sqrt{2\alpha}} (1 - e^{-2\alpha l})^{1/2}$$

The generation current $2i$ is found from the cage measurement of the radio noise excitation function Γ_c for a particular conductor or multi-conductor bundle. Γ_c is measured on a quasi-peak basis at a given centre frequency (500 kHz, Bandwidth 9 kHz) and conductor surface gradient (1, 3).

The relationship between conductor surface gradient and the excitation function is shown for a particular practical case in figure 2(1,2).

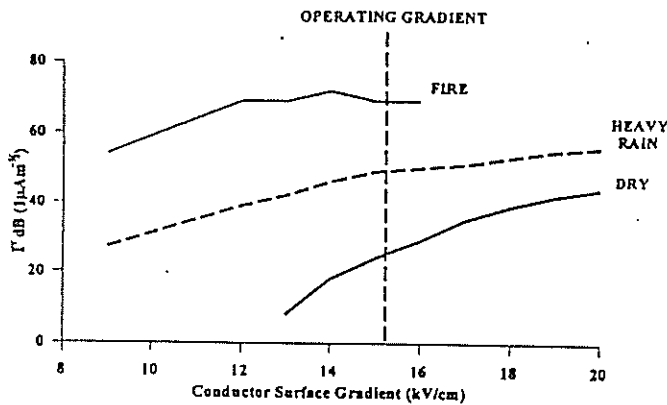


Figure 2 Excitation functions for a twin Dinosaur conductor bundle

To transform the cage measurement of Γ_c to the new generation current level $2i'$ applicable to a defined line geometry, the following relationship is used (3, 5):

$$2i' = \Gamma_c \frac{C}{2\pi\epsilon_0} \quad (2)$$

Here C is the capacitance between the given conductor and a specified earth reference, and ϵ_0 is the permittivity of free space. Substitution in (1) gives:

$$I' = \frac{1}{2\sqrt{2\alpha}} (1 - e^{-2\alpha l})^{1/2} \Gamma_c \frac{C}{2\pi\epsilon_0} \quad (3)$$

For a simple single phase geometry (conductor of radius r at height h above an infinite earth plane). Thus,

$$I' = \frac{1}{2\sqrt{2\alpha}} (1 - e^{-2\alpha l})^{1/2} \Gamma_c \frac{1}{\ln \frac{2h}{r}} \quad (4)$$

The above equation is used to calculate the total propagated current I received at the termination A as a function of line length.

In applying equation (3) to wet or heavy rain conductor corona, it is assumed that the rain occurs uniformly along the line, away from the termination point. This will be a more severe case than the sporadic distribution of zones of rain along a line.

3.2 Fire Noise

Unlike conductor corona noise, fire noise is modelled as a point source of injected current at some point x along the line. The noise current then propagates in both directions, as shown in figure 1.

As regards the variation in quasi-peak level with measurement frequency for both fire and wet conductor corona noise, a reduction in level of 5 dB over the frequency range 200 to 500 kHz has been found experimentally (2).

3.3 Polluted Insulator Noise

The cumulative interference from polluted insulators is produced by a series of discrete point sources as shown in figure 1.

The total propagated pollution current I_i received by the line terminations is derived from the quadratic sum of the individual attenuated currents. Thus,

$$I_i = i_{i1} \sqrt{\sum_{k=0}^n (e^{-\alpha ks})^2} \quad (5)$$

where s = separation between point sources (ie span length)

k = number of spans

and $2i_i$ = injected quasi-peak current (in a specified frequency band) for each insulator (vee strings not considered).

For ease of analysis, it is assumed that the insulators are equi-distantly spaced, and that the injected currents are the same. Uncorrelated noise currents are also assumed.

4. PRACTICAL CASE

The foregoing equations do not lend themselves to purely analytical study, and have therefore been applied to a

specific practical case. Figure 3 shows the total noise currents, as a function line length, as received at the termination A in figure 1, for a centre frequency of 50 kHz.

The conductor surface gradient and attenuation parameters have been derived from a twin "Dinosaur" bundle as used on many Eskom 400kV lines. The conductor corona and fire noise data has been taken from (1) and (2).

The likely maximum quasi-peak noise levels at 50 kHz generated by polluted porcelain longrod insulators have been derived from laboratory work done by Cunliffe (4). An amplitude range of 80 to 100dB ($1 \mu\text{V}/300\Omega$) has been considered; this refers to conditions of severe pollution.

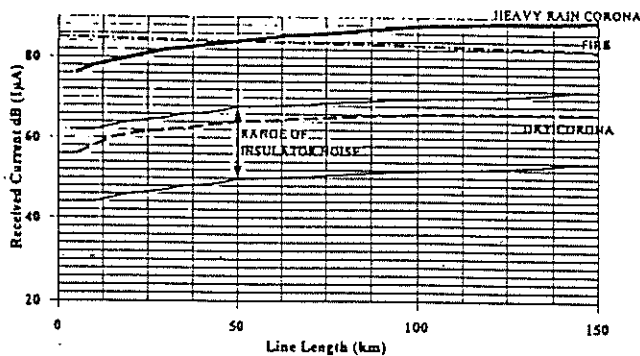


Figure 3 Propagation at 50 kHz on the twin dinosaur bundle.

5. TENTATIVE FINDINGS

The tentative findings from the frequency-domain quasi-peak method, as applied to the twin dinosaur conductor bundle, are :-

5.1 Fire noise should be distinguishable from **dry conductor corona noise** for line lengths of at least 200km.

5.2 If the first few tens of kilometres of line away from the measurement point (substation) are exposed to heavy rain, detection of fire noise will not be possible, unless the fire is 50km or less away from the substation.

5.3 The noise from polluted insulators in general will not exceed that of dry conductor corona. This suggests that it might not be feasible to use the above frequency domain method for the reliable identification of high levels of pollution noise on a 400kV line, at least.

5.4 The fire/wet corona cross-over distances at 50 and 500 kHz are similar; this shows that the choice of measurement frequency is not critical.

5.5 The results show that an amplitude discrimination threshold between fire and heavy rain noise (quasi-peak) can be rationally set. Its magnitude will depend on the conductor bundle, line voltage, line length and local weather patterns.

5.6 Although the numerical example deals with a particular case, a simple methodology has been outlined. This will allow alternative conductor bundles to be studied.

Rigorous modelling of modal propagation will lead to slightly higher received noise levels arising from conductor corona activity in the first few kilometres away from a substation. Beyond this range, the low loss modal component will predominate, leading to results similar to that in figure 3.

6. CONCLUSIONS

6.1 A simple method of transforming quasi-peak noise levels, as measured in a corona cage, to the corresponding levels on an elementary single phase overhead line has been described.

6.2 Simple models for describing the propagation of noise from polluted insulators, fires and conductors in corona have been developed. This makes it possible to study the differences between the various noise levels as a function of line length.

6.3 The results, for a specific case, suggest that in ultimate practical detection systems the following findings apply :

- (a) Discrimination thresholds can be rationally set.
- (b) Fire noise generated beyond a few tens of kilometres from a substation is not detectable above that of heavy rain corona noise with quasi peak methods.
- (c) Noise from severely polluted insulators will generally not be detectable above that of conductor corona in dry conditions.
- (d) More sophisticated pattern recognition techniques have to be used to overcome the above limitations with quasi-peak measurements.

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INSULATOR POLLUTION PERFORMANCE TEST STATION: DESIGN AND OPERATION

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ABSTRACT

A marine natural ageing insulator pollution test station has been established at Koeberg Nuclear Power Station near Cape Town.

Koeberg Insulator Pollution Test Station (KIPTS), will be used by Eskom to evaluate the performance of non-ceramic insulators and coatings, energised at 22 and 66 kV. As a spin-off valuable research data will be obtained.

The design principles and operation of the test station are presented in this paper.

INTRODUCTION

Glass and porcelain (ceramic materials) have dominated the insulator market for many years. The performance of this type of insulator has been the subject of many research projects and application principles are well established [1].

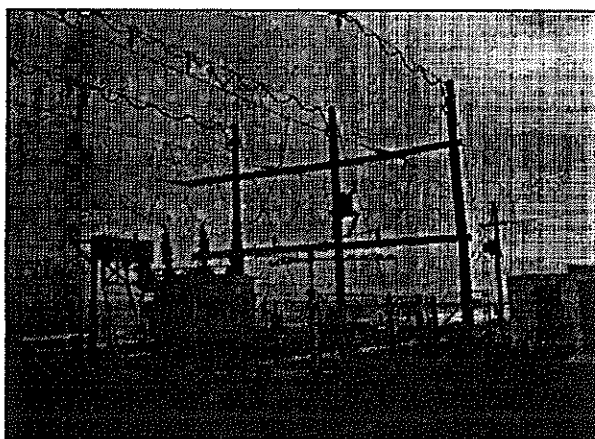


Figure 1: Photo of KIPTS. Note insulators not fitted at this stage.

However, developments in material science technology has led to new insulator designs, using polymeric (non-ceramic) materials. These include cyclo-aliphatic epoxy resin cast insulators and composite insulators, consisting of a glass fibre core and polymeric sheds. These materials are often hydrophobic as opposed to hydrophilic ceramic insulators. As a result of their low wettability, these insulators perform exceptionally well in extremely polluted environments. During prolonged exposure to heavy pollution, the

hydrophobicity deteriorates, resulting in higher leakage currents and further degradation. The hydrophobicity can however recover during subsequent periods of low pollution and moisture [2].

New materials and insulator designs regularly become available on the market and incorrect choice of material or insulator may prove costly in later years. International research on insulator testing and ageing has led to basic guidelines on the assessment and application of non-ceramic insulators [3]. However, environmental conditions along the Cape west coast are somewhat unique and it was therefore decided to build a marine natural ageing insulator test station at Koeberg Nuclear Power Station. It is foreseen that monitoring of the insulator performance over a period as short as one year may give a useful indication of the relative performance of insulators.

KOEBERG INSULATOR POLLUTION TEST SITE (KIPTS)

The test site is situated along the west coast, 50 meters from the sea on property made available by Koeberg Nuclear Power Station. An area of 1050 m² is used to house the 66 kV and 22 kV insulator test bays complete with control room, environmental monitoring station and logger system.

The climate at KIPTS is characterised by dry summers, high winds, mist banks and heavy marine pollution. The rainfall occurs mainly in winter. The pollution index at KIPTS is of the order of 2000 [4] which is extremely high and would be classified as 'Very Heavy'. Therefore KIPTS constitutes the ideal marine pollution area in which to evaluate insulators and insulator materials.

GENERAL DESIGN

A photograph of the test site is shown in figure 1. The 66 kV test transformer and busbars, control room and the mobile 22 kV test unit can be seen.

The insulator test station was designed to test 30 x 66 kV and 15 x 22 kV insulators simultaneously. A minimum specific leakage path of 25 mm/kV was chosen. The longitudinal electric stress (kV/m) is kept similar between the various test insulators for comparison purposes. All the different coatings to be

tested are applied to the same type of porcelain longrod for comparison purposes.

A portable hydraulic hoist (2m x 1m platform, lifting capacity of 500 kg) was specially designed to install and inspect the test insulators. A limit switch which trips the station HV supply was installed on the garage door where the hoist is stored to prevent its operation when the station is energised. A single line diagram of the electrical layout of KIPTS can be seen in figure 2.

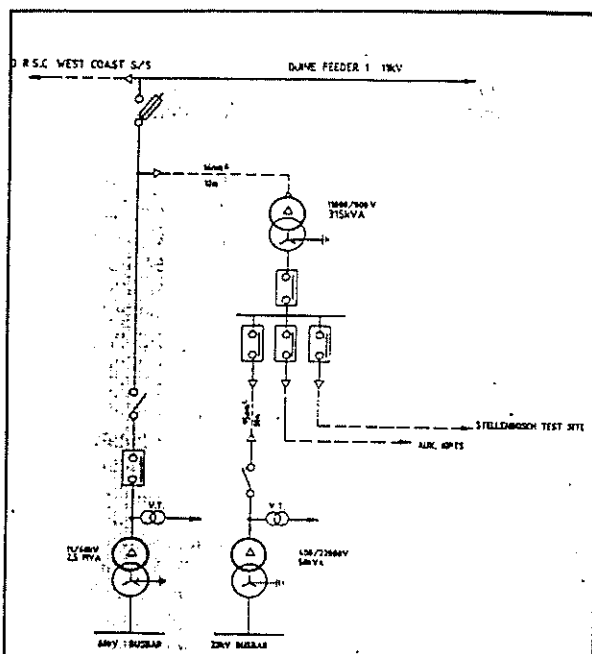


Figure 2: Electrical Schematic diagram

66 KV TEST BAY

The 30 x 66 kV test insulators, 10 per phase, are energised from a 11 kV supply stepped up to 66 kV via an 11/66 kV, 2.5 MVA delta/star transformer. An explosive Mace fuse (0.8 A) is used to isolate the faulty insulator from the 66 kV busbar in the event of a flashover. The system is suitably protected and controlled from the control room. An interlock on the access door of the chamber housing the transformer trips the supply when opened. The transformer has one extra non-standard protection feature: when the leakage current over its bushings exceeds a predetermined level it will indicate and then trip its supply. The star point of the transformer is directly earthed to the station earth.

The short circuit current (I_{sc}), phase to earth is 217 A (rms) at the 66 kV busbar. The maximum peak current for an insulator with creepage of 25 mm/kV is given as 1.35 A (peak) [5] thus I_h (peak) for 10 insulators is 13.5 Amps (peak). This leaves us with an I_{sc}/I_h ratio of 16.1 which is larger than the recommended [5] ratio of 11.

The 11 kV supply to the transformer is monitored by

the logger system via a phase to phase VT. Using the measured transformer ratio the voltage on the 66 kV bus can be determined.

22 KV MOBILE TEST UNIT

The 15 x 22 kV test insulators, 5 per phase, are energised from a 400 V supply stepped up to 22 kV via a 400 V/22 kV, 50 kVA delta/star transformer. An explosive Mace fuse (0.8 A) is used to isolate the test insulator from the 22 kV busbar. The system is suitably protected and controlled from a control box attached to the unit. A red light beacon indicates that the unit is energised. An interlock on the access door of the chamber housing the transformer trips the supply when opened.

The star point of the transformer is directly earthed to the station earth.

The short circuit current, I_{sc} phase to earth is 86 A (rms) at the 22 kV busbar. The maximum peak current for an insulator with creepage of 25 mm/kV is given as 1.35 A (peak) [5] thus I_h (peak) for 5 insulators is 6.75 A (peak). This leaves us with an I_{sc}/I_h ratio of 12.7 which is larger than the recommended [5] ratio of 11.

The 400 V supply to the transformer is monitored by the logger system via a phase to phase VT. Using the measured transformer ratio the voltage on the 22 kV bus can be determined.

LEAKAGE CURRENT MONITORING

The leakage current flowing over the various insulators and coatings energised at 22 and 66 kV are monitored. This information is used to assess the insulators' performance.

The following leakage current parameters are monitored:

- * Maximum positive and negative peaks of the leakage current over a set interval (Highest peak current, I_h).
- * Sum of the positive and negative leakage current waveform samples over a set interval (Charge measurements, Coulomb).
- * Sum of the square of the leakage current waveform samples over a set interval (Coulomb-Ampere).
- * Counters (4 Pos and 4 Neg), counting the number of times the leakage current peaks exceed programmable thresholds over a set interval (Surge counting).
- * Waveform of the maximum leakage current (100 ms period) over the insulator or coating each day.

- * Waveform of the leakage current (100 ms period) if a preset threshold is exceeded.
- * The time-to-flashover for each insulator or coating.

The IH48, a dedicated transputer based, 48 channel logger and sensor system, was specially developed to achieve the above.

The IH48 hardware has the following characteristics:

- * Records all the above data for up to 48 channels simultaneously and stores it to a 4 M-byte non-volatile on-board memory.
- * The sample interval can be set between 1 min to 1 hour.
- * Very low impedance galvanically isolated current sensors which can measure 0 to 3 A (rms) are used. The 0 to 40 mA (rms) sensor outputs, are connected via shielded twisted pairs to the logger.
- * High accuracy of below 1% (12 Bit), with a sampling rate of 1 kHz per channel is achieved.
- * The logger can be used as digital scope to take real-time snapshots (100 ms) of the current waveform for a selected channel.

The IH48 software has the following characteristics:

- * Fully menu driven with on-line help function.
- * Works as stand alone unit. Data (1.5 Mega-byte) is automatically down-loaded to a hard-drive daily.
- * Auto restarts after power failure.
- * The logger is configured from within the database connecting types of insulators or coatings to channels.
- * Software calibration on input channels.
- * Integrated database with information concerning manufacturers and their different insulators or coatings. Database stores and manages the recorded data. History of manufacturers and insulators with full editor and clipboard functions.
- * Graphical interface for viewing recorded data. The graphic features are: primary and secondary axes, legends, line styles and colours, headings, grids, labels, zoom function, cross-hair and print-graph facility.
- * Displays statistical data of the current surge counters and times to flashover.

ENVIRONMENTAL MONITORING

The following parameters are monitored:

- * Wind speed (m/s) and direction (deg)
- * Relative humidity (%)

- * Temperature (deg C)
- * Rain (mm)
- * UV Solar radiation (300 - 400 nm) (mW/m²)

The SOLUS, an environmental monitoring and data logger system, is used to monitor the above parameters. The data is continuously stored in non-volatile memory and then down-loaded to a hard-drive.

Equivalent Salt Deposit Density (ESDD), Non Soluble Deposit Density (NSDD) and Dust Deposit Gauge (DDG) measurements are taken on a monthly basis. This data will be used in conjunction with the Insulator Pollution Monitoring Apparatus (IPMA) [6] to monitor the test site pollution severity.

Directly sampling a known volume of air through a glass fibre filter and then analysing the sample is under investigation. This will also be installed at KIPTS.

INSULATOR POLLUTION MONITORING APPARATUS (IPMA)

An automatic insulator pollution monitoring apparatus (IPMA) was designed and built to measure the surface conductance of test insulators. It's design is based on the CESI pollution monitoring equipment (PME) [7] and is an extension of the unit previously built by the University of Stellenbosch [8].

Its sole purpose is to measure the pollution layer conductivity on the surface of the test insulator without destroying the pollution present. This measurement will be used to determine the pollution severity of the surrounding environment.

The IPMA is currently being calibrated in the laboratory at Stellenbosch University before installation at KIPTS. Calibration curves of Surface Conductivity (uS) to ESDD (mg/cm²) to Clean-fog (Leakage current in mA) to Pollution Index (uS/cm) are being set up, normalised to a 30 day month.

IMAGE INTENSIFIED VIDEO AND DATA CAPTURE SYSTEM

An image intensified video camera (COROCAM), highly sensitive in the ultra violet spectrum, has been locally developed by the CSIR and the University of Stellenbosch to look at corona discharges. This is a valuable tool in the field of insulator pollution in that it can pin point the exact location of minute discharges on the insulator.

A frame grabber card has been developed to capture eight consecutive video frames. A synchronisation circuit used with the frame grabber card, in conjunction with a high speed (3 MHz) data capture card or digital storage scope, enables us to capture the

leakage current waveform over the insulator in synchronism with the video recording of the discharge.

The above system is used at KIPTS to record the discharges and leakage current waveforms over the test insulators on a regular basis. The information is kept on record to help with the performance assessment of the various insulators and coatings under test.

MATERIAL ASSESSMENT

A full assessment of the insulator material will be performed on a regular basis.

On site we are currently looking at hydrophobicity using the Swedish system [9]. Video recordings are made of the insulator surface on a regular basis.

CONCLUSIONS

KIPTS will be used as a decision and research tool in the field of insulator pollution.

The information obtained will aid in the setting up of an accelerated ageing chamber in South Africa.

Eskom will make more cost effective decisions regarding outdoor Insulators in the future for it will be based on scientific facts obtained from our insulator research program.

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WHATEVER HAPPENED TO PLAIN ORDINARY KILOWATT-HOURS?

NEW INTERRUPTIBLE & CURTAILABLE TARIFFS

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Eskom

ABSTRACT

This paper outlines basic premises for the design of interruptible and curtailable electricity tariffs used by many world-wide utilities to offer reduced rates to customers in return for load that can be taken off the system (partially or completely) during peak periods or system emergencies.

INTRODUCTION

As recently as the late '80's electricity offerings in South Africa were still relatively unexciting. Kilowatt-hours were packed in plain ordinary packages. But today, with the advent of Time-of-Use tariffs, many kilowatt-hours are individually packaged in significantly coloured wrappers. Red for peak; green for off-peak and yellow for standard kilowatt-hours.

Consequently many customers have an opportunity to determine what kind and what mix of kilowatt-hours they purchase.

Even more kinds of kilowatt-hours are expected to be on sale in the 90's. And not only kilowatt-hours! Services generally are expected to be considerably more varied with increasing emphasis on industry sponsored demand-side management programmes to:

- (a) meet the challenge of supplying today's customers with electricity at lowest possible prices and
- (b) to off-set the risk of prematurely committing to overly expensive long term capacity investments to meet the peak demand requirements of tomorrow's customers

Some of the most effective demand-side measures for addressing these challenges and risks are interruptible and curtailable tariffs (1)

TARIFF DEFINITIONS

Interruptible tariffs offer participants a price reduction in return for allowing part of their load to be interrupted by the utility (usually with little or no advanced notice, by

means of a remote signal from the national or regional control centre).

Curtailable tariffs offer price reductions in return for voluntary load curtailment on request from the utility. The request is usually made several hours in advance of when the curtailment is needed.

Unlike interruptible tariffs, where the decision to interrupt is normally out of the customers hands, with curtailable tariffs customers decisions to curtail load are normally influenced by the high price(s) prevailing during the curtailment period.

WHY NOW?

As was the case with Time-of-Use tariffs it is not appropriate to wait until capacity shortages start to be felt before developing interruptible or curtailable tariffs. It requires time to develop rates, canvass prospective customers and familiarise them with the basic principles. By the time capacity starts to run out it would be too late to nurture interruptible or curtailable loads in time to benefit long term capacity planning.

BENEFITS

Interruptible and curtailable service programmes offer reciprocal benefits to customers and utilities supplying them.

The primary benefit to customers is an opportunity to reduce electricity costs by rescheduling industrial/ commercial processes or household loads.

In particular these programmes offer customers having their own back-up generation, an opportunity to reduce electricity purchases when utility supply costs are highest.

From a utility viewpoint, offering price reductions to meet peak load requirements can often be cheaper than operating existing peaking capacity or investing in new capacity.

Another benefit is the deferral of transmission and distribution investments. Areas on the network with high growth patterns would need to be targetted to derive maximum

benefit. Customers with back-up generation could contribute greatly to localised system security, enhancing the utility's flexibility in dealing with dynamic power disturbances.

Finally, whilst customers have grown to expect a high degree of reliability from reserve margins based on theoretical loss of load probability calculations, the cost of carrying large reserves could be more than it is worth to them. Offering reduced prices for lower service reliability would make it possible to re-determine generating reserve requirements using economic value-of-service criteria instead of probability statistics.

CONSTRAINTS

From a customer perspective, tariff benefits need to be weighed against the cost of an outage, and if applicable, the cost of installing and operating back-up generation. When hours of operation are low, as they typically are, costs of back-up generation can be significant compared to the cost of purchases from the utility.

From a utility perspective, benefits need to be weighed against the cost of incentives, programme costs and revenue lost during interruptions, (albeit typically small due to the limited hours of operation).

SIGNALLING INTERRUPTIBLE/ CURTAILABLE EVENTS

With a no notice interruptible tariff load shedding is achieved automatically via utility activated remote controlled relays on the customer's premises.

Under curtailable tariffs prospective curtailment opportunities are signalled via electronic or telecommunication methods. Depending on the type of tariff, prevailing rates are also confirmed. Rate structures may vary from a single very high "super-peak" rate to a sequence of rates applicable to the full peak day.

CONVERTING COSTS INTO PRICE INCENTIVES

Because peak load reduction is the primary motivation for interruptible and curtailable programmes, avoided peaking generation costs are typically the starting point for setting price incentives.

Finalising price incentives is, however, a qualitative, judgemental process. Appropriate allocation of generation and transmission & distribution costs is not a generic question but rather it is specific to each participant. Participants will have different load characteristics and while it may be beneficial to interrupt (or request curtailment of) the loads of some customers, it may not be beneficial in cases where, for example, on-peak load factors are low.

Equity dictates that price incentives should be in proportion to customers' on-peak load factors to reflect the probability that a significant load will be available at the time of a

curtailment request. Furthermore billing credits would need to be based on interruptible load nominated during system peak periods, because the incidence of interruptions or requested curtailments would be considerably less at other times.

CUSTOMER OUTAGE COSTS

Depending on circumstances customer outage costs may need to be factored into price incentives.

Customers don't really care what causes an outage. They view reliability against a range of parameters such as:

- Duration of outage
- Frequency of outage
- Timing of outage (hours of the day, day of the week, season of the year)
- Extent and timing of notice
- Magnitude, extent and persistence of curtailments

Customers are affected differently. Continuous processes experience more disruption than batch processes because there is less likelihood of slack in the operation and a greater likelihood of an irrecoverable loss of production. Thermal-electric processes such as refrigeration, baking, metal or glass melting may also experience significant product loss during an outage.

PENALTIES FOR NON-COMPLIANCE

Of 50 interruptible programmes surveyed by R Kovach (2), two thirds had some form of failure to interrupt penalty. Of these the most common form of penalty was an additional charge per kW/kVA of demand. Five programmes incorporated demand charges amounting to between 5 and 10 times the normal rate and eleven programmes contained a provision for terminating service if customers failed to interrupt when requested to do so.

METERING & BILLING IMPLICATIONS

In tariff structures incorporating penalties, billing debits for non-compliance are settled after the event but credits for compliance can be settled before or after the event.

Metering and billing are likely to present more challenges where accounts are settled after the event because the load curtailed (or not curtailed) must be determined, translated into billing units, and featured in the bill. Under these circumstances metering algorithms may be required to estimate the appropriate load and enhancements to existing billing systems will be necessary.

INTERNATIONAL PRACTICE

Electricite de France (3)

Electricite de France offers a Peak Day Demand Reduction option to all its customers from high voltage industrial to low voltage residential. The peak period spans 18 hours per day for 22 days in the year. Peak days are determined in real time by the utility who give customers at least 30 minutes notice. Prices during other days are as specified in standard tariffs.

Not only is this option attractive to customers who are able to reschedule industrial/commercial processes or household loads it is also attractive to customers having alternative energy supplies. By using their alternative supplies on peak days customers can reduce costly purchases from the utility.

A refined version of the Peak Day Demand Reduction option (called the Modulatable option) is offered to customers at medium voltages and above. In this version, the year is divided into three types of week (24 low season weeks, 19 mid-season and 9 winter season) each having their own characteristic rate periods and price levels. The weeks are not necessarily chronological. Customers are notified of the week type by the utility as the year progresses. Peak days may be signalled at anytime but usually they occur during the 9 movable winter weeks.

Electricite de France introduced the modulatable version to encourage the use of electric arc heating on days when supply costs were lowest, but it is also popular with industrial customers who can manage their load without the aid of dual energy systems.

These options have become very popular, particularly with large customers as shown in the following table.

Niagara Mohawk (4)

Niagara Mohawk Power Company, Syracuse, New York State, offers customers a Voluntary Interruptible Pricing Programme to reduce load when system emergencies or power system economics dictate. There are three options under this programme namely Day Ahead Notice, Same Day Notice and Short Notice (10 mins). Each offers participants an up-front incentive per kW of nominated interruptible or curtailable load. Participants must have demands in excess of 500 kW. The minimum nominated curtailable load is 100 kW and the maximum duration of a curtailment is 8 hours per day.

Day Ahead participants are notified the previous day not later than 16h00. The rates applied during the curtailment period are day ahead expected hourly marginal energy prices for each hour of the curtailment. Curtailment down to an elected firm power level is voluntary and any load taken in excess of this level is charged at hourly marginal energy prices.

The Same Day option specifies a minimum of 4 hours notice. If load is not curtailed down to an elected firm power level a predetermined c/kWh penalty is levied.

The Short Notice option is designed for emergency situations only and the incentive offered is the lowest of the three because of the low probability of an occurrence. Unlike the Same Day and Day Ahead options where an assessment is made of the average load available at customers facilities during the exposure period, a history of the participant's 30 minute metered demands is used as the basis for interruptible load nominations. A c/kWh penalty is also levied for non-compliance.

In 1990 Niagara Mowhawk had seven customers on the scheme and load relief at peak times was 8MW.

Size	Voltage level	NUMBER OF CUSTOMERS		
		Total	PDDR	Modulatable
≥ 40 MW	EHV	90	34	12
10 MW - 40 MW	HV	320	80	41
250 kW - 10 MW	MV	130 000	3 100	100
36 kVA - 250 kVA	LV	120 000	1 500	-
≤ 36 kVA	LV	26 000 000	400 000	-

LOAD CURTAILMENT AUDITS

A survey carried out by D P Theron (5) estimates some 4000 Megawatts of potential interruptible and curtailable load in South Africa presently comprising 700 MW of municipal ripple/radio control, 300 MW of available gas turbine capacity belonging to customers (back-up generation), and 3000 MW of industrial load.

In order to further assess the potential for interruptible/curtailable load at grass roots level, detailed audits are expected to start next year.

The objectives of this type of audit are to:

- identify and understand critical loads in terms of possible shut down, load shifting or substitution options
- obtain information on how different equipment operates during system peak periods
- identify, quantify and discuss load curtailment opportunities with prospective participants.

CONCLUSION

Whilst the exact structure of the new electricity industry remains uncertain one thing is definitely clear. Kilowatt-hours will become available in more exciting packages.

Interruptible and curtailable services are widely known as low price/low reliability options but they can also become a low price/high reliability option when combined with customers' own back-up generation.

In view of these benefits, a variety of experimental options embodying the principles and practices discussed in this paper, are being considered for implementation in 1994.

These are likely to comprise a no notice interruptible option; a same day/day ahead notice curtailable option (similar in design to the Electricite de France "Peak Day Demand Reduction" tariff); and a combination of the two.

Thus yet another kind of kilowatt-hour, the kind used (or not used!) during notified curtailment periods, is in future likely to be the crimson coloured kilowatt-hour, signalling very high supply costs in excess of red Time-of-Use prices.

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POWER FACTOR CORRECTION AND ENERGY COSTS SAVINGS

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ABSTRACT

Industrial and other large energy consumers are encouraged by utility companies to operate at higher power factors, to minimize system power losses and voltage drops, and to increase available system MW capacity. Consumers also stand to gain in terms of reduced energy costs. A user friendly FORTRAN software for evaluating optimum power factors for any operating conditions has been developed. The maximum savings, of the consumers studied, varied between 5% and 20%. Possible savings can be increased if utility companies made it even more attractive for consumers to operate at higher power factors.

1. INTRODUCTION

Utility companies apply special tariffs to industrial and other large consumers of electricity, to encourage them to operate at higher power factors. The tariff usually consists of a charge on a maximum demand and a power factor penalty, in addition to the charge on energy (kWh) consumed.

It is to the advantage of utility companies for consumers to operate at high power factors because then there is improved voltage regulation and system operates more efficiently. On the other hand for a given kVA, there is extra MW capacity at higher power factors, and so the utility companies can afford to connect more loads at no extra investment costs.

To the consumer, power factor correction offers economic advantage in terms of savings on energy bills by reducing the maximum kVA demand and avoiding the penalty on power

factor. Installation of power factor correction equipment requires capital investment, and hence annual capital costs. Therefore, for each consumer, there is a power factor at which the economic benefits of installing the equipment (i.e. saving in energy costs) equals the cost of the capacitors. This is known as the optimum power factor.

In this study, the optimum power factors for a number of consumers are evaluated using a specially developed computer programme. The power factor is calculated by taking into account the tariff structure and interest rates on capital. This study is part of a wider one undertaken to assess the load consumption characteristics of different types of consumers.

2. DATA ACQUISITION

Data on energy consumption was obtained from industries (large, medium and small) around Nairobi and from the University of Nairobi. The data from industries was obtained by means of a questionnaire which sought to know the monthly maximum kVA demand, energy (kWh) consumed, the average power factor, and number of operating hours per month. Data from the University was obtained by making instantaneous kW and kVA readings at hourly intervals, daily, for about 4 months.

3. DATA EVALUATION

The data was analysed using a specially developed, user friendly FORTRAN computer program (1). It calculates the optimum power factor by taking into account the tariff structure, and the capital charges (i.e. interest and depreciation). The program also evaluates the cost of the required power factor correction

equipment, the percentage savings and the payback period. The operator of the program is prompted to input the monthly energy consumption (kWh) and the number of operating hour (or just the monthly kW demand), the maximum kVA demand, the average power factor and the cost per kVAR of power factor correction equipment. The operator has also to specify the category of the load, as specified in the Byelaws of the utility company, and the interest rate chargeable on capital investments. The savings on maximum kVA demand and power factor penalty were calculated at different values of power factor, increased by 1%, from the operating level up to 1.00. Although the energy (kWh) consumed is independent of the operating power factor, the charge on the kWh is increased by 1% or 2% (depending on the load category) for each complete 1% decrease in power factor below the minimum recommended level (i.e. 0.9) (2)

A summary of data and results obtained for each of the consumers studied is shown in Table 1. From the data analysis it was also possible to obtain plots showing variations of savings with power factor as well as the dependence of optimum power factor on the rate of interest. Typical plots are shown in Figures 1 and 2.

4. DISCUSSION

Theoretically, the optimum power factor is obtained by evaluating

$$\frac{d[Qf(\theta)]}{d\theta} = 0$$

where Q is the net savings on energy costs, and θ is the angle corresponding to the improved power factor. Practically this is not possible to evaluate because of the power factor penalty clause in the tariff structure. It is, therefore, necessary to use the step by step method of calculating savings at different power factors. The optimum power factor is that which gives maximum savings. For power factors below the optimum value, the savings decrease due to increased penalty charges, and at higher values of power factor, the savings decrease

because of increased charges on power factor correction equipment (Figure 1).

Maximum savings varied from 5% to about 20%, depending on the initial power factor and the chargeable rates of interest. There was a direct correlation between the change in power factor (i.e. from the initial value to the optimum value) and the maximum savings. For example Firm C with a 36% change in power factor, had a maximum saving of 21%, whereas Firm A had a saving of about 10% corresponding to an 11% change in power factor.

The optimum power factor was dependent on the rate of interest on the capital investment. It decreased with interest rate down to a minimum level of about 0.9 which remained constant with interest rate for values higher than 10% (Figure 2).

A number of consumers studied were not aware of the concept of optimum power factor. Many were content with operating at the minimum recommended level to avoid the penalties. This apparent lack of interest is probably because the accrued savings were considered small. This attitude would have been different if the utility company imposed more stiff penalties (because then the the Savings would be substantial even for small changes in power factor), or if it introduced a system rebates, (i.e. to reward consumers operating above the minimum recommended power factors).

5. CONCLUSIONS

In this study it has been possible to develop a simple user friendly, software, which can be used to calculate the optimum power factor by taking into consideration the tariff structure, technical and economical operating conditions.

Since the utility companies stand to gain more if consumers operate at higher power factors, then there is need for them to make it more attractive for consumers to operate at increased power factor levels for example by introducing rebates, in addition to more stiff penalties, or raising the minimum required power factor level.

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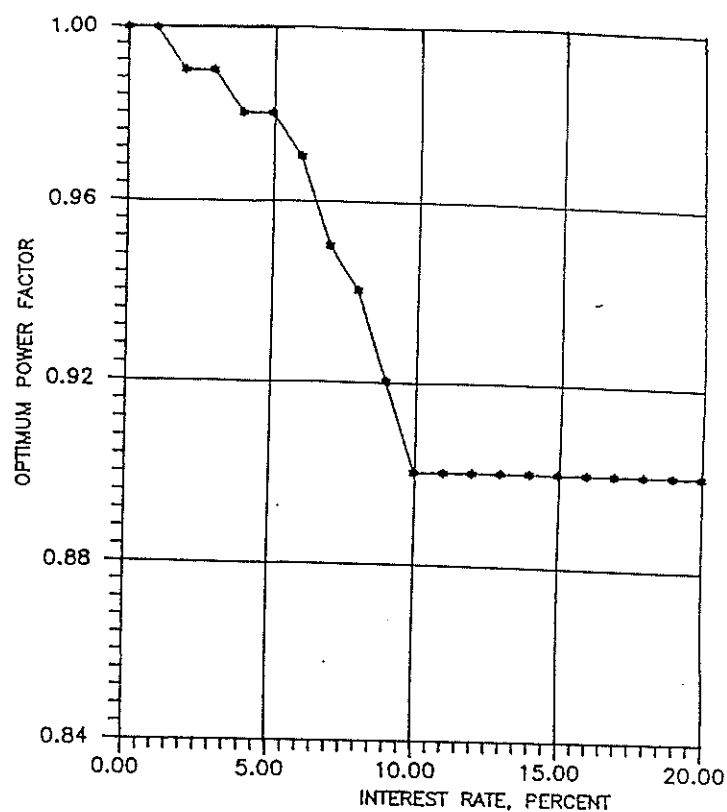


Figure 2: Variation of optimum power factor with interest for Firm B.

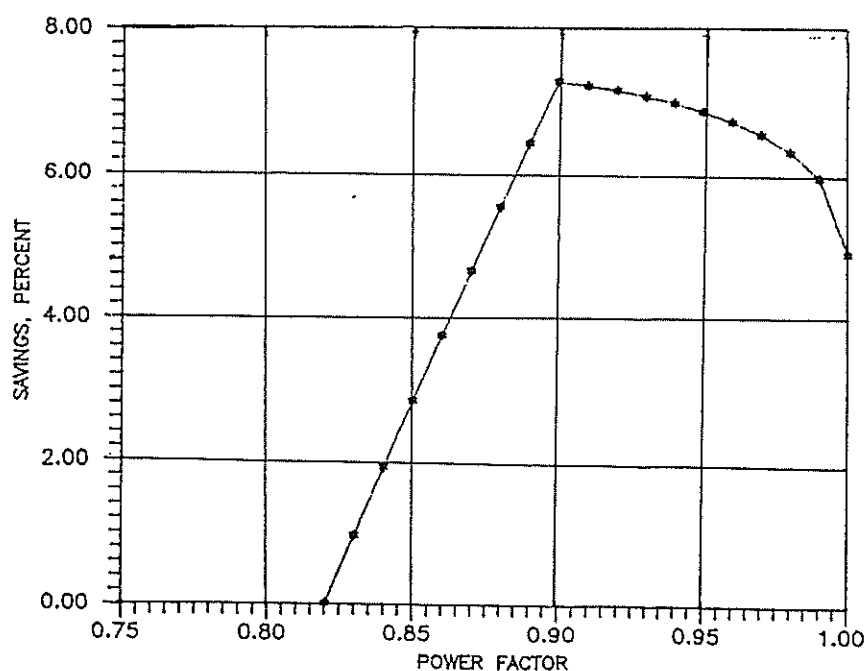


Figure 1: Variation of savings with power factor for Firm B.

Table 1: Summary of Results

Consumer	Average KW Demand	Max. KVA Demand	Initial Pf	Optimum Pf	Max. Savings (%)	Inter. Rate (%)
Firm A	4730.0	5330.0	0.79	0.90	9.9	12.5
Firm B	220.0	303.0	0.72	0.90	7.4	10.5
Firm C	2105.0	3104.0	0.64	0.90	21.0	9.5
Firm D	802.0	671.0	0.76	0.90	13.0	8.5
Halls	86.5	107.6	0.80	0.90	9.0	10.0
Adm. Blocks	93.0	120.8	0.85	0.90	5.0	10.0
Eng. Block	51.0	60.1	0.84	0.90	8.0	10.0
Parklands Campus	23.8	28.7	0.83	0.90	8.0	10.0
Chiromo Campus	190.0	205.6	0.92	0.93	5.0	10.0

COMPUTER INTERFACE FOR THERMAL ENERGY METERS

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1. ABSTRACT

Thermal energy meters are used to measure the kilowatt hour, voltamp and maximum demand from mainly factories and businesses where the energy consumptions are high. A need existed to reduce the reading time of the thermal energy meters and read the meters more accurately, to ensure that the energy account reaches the client sooner.

An electronic meter could be used for this purpose but would not justify the costs. No substitute product could be found, and the interface was developed. The interface consist of hardware and software with possibilities for future development.

2. INTRODUCTION

The purpose of the research is to analyse the Sangamo 1000VA and the Sangamo 2000VA electro-mechanical analogue revenue energy meters i.t.o. the VOLT-AMP-, maximum demand- and kilowatt hour (kWh) readings, in order to identify the key success factors that will ensure that the mentioned readings, from the energy meters, can be registered and stored even when a power failure occurs, and can be dumped to a personal computer.

3. THE PROJECT

3.1 Delimitations

The study was limited to power meters that would read the VOLTAMP-, maximum demand- and kilowatt hour readings of businesses and industrial premises. The emphasis of the study was to construct and control an interface capable of operating between a thermal energy meter and a computer via a transmission medium.

3.2 ERRORS

All electronic components or instruments have, to some extend, an error or inaccuracy. It is important to understand how they combine to create even greater errors in measurement systems. For this project accuracy is extremely important and therefore the worst combination of errors that could effect the different meter readings was considered. It is therefore clear that a good understanding of the circuit operation of the meter is of great importance.

3.3 INTERFACE POSISTION

For the following reasons, special attention was given to the positioning of the installation of the computer interface:

- (a) Power supply must be available.

- (b) Access must be easy for maintenance etc..
- (c) Enough space must be available for the installation of the data transmission medium in the direct vicinity.

3.3.1 INSIDE THE METER

Due to the following reasons, the installation cannot be done inside the meter:

- (a) The inside space is very limited, only one of the boards could have been designed to fit inside which would have resulted in excess wiring linking the two boards.
- (b) Installation and maintenance would have been complicated.
- (c) Not cost effective.

3.3.2 CLOSE TO THE METER

The most logical position for installing the interface is as close as possible to the energy meter whose readings has to be converted. Thus any suitable medium can be used to link the interface to the computer.

3.4 HARDWARE

3.4.1 MINIMUM HARDWARE

The following hardware is the minimum hardware necessary to construct the Computer Interface.

- (a) An 8-bit microcontroller.
- (b) A 10-bit resolution analog-to-digital converter.
- (c) Power supply from one phase.
- (d) A signal from the remaining phases.
- (e) Standby power supply in case of power failure.
- (f) Read-and-Write memory

- (RAM) for data storage.
- (g) Erasable-Programmable memory (EPROM) for program storage.
- (h) Provision for Press-to-Talk (PTT) and Carrier Detect (CD) for radio link.
- (i) Receive (RX) and Transmit (TX) for communication.
- (j) The signals to be converted from the meter.

As a result of the preparatory study it was decided to divide the hardware into the following two boards:

- (a) The power supply and I/O board.
- (b) The processor board.

3.4.2 PROCESSOR BOARD

3.4.2.1 NECESSITY

Because it would be difficult to control the inputs, outputs, transmission medium and the communication, it is necessary to use a microprocessor environment.

3.4.2.2 LAYOUT

- (a) The microprocessor
- (b) The latch
- (c) The data memory
- (d) The program memory
- (e) Standby for power failure
- (f) Receive/Transmit for communication
- (g) Real Time Clock

3.4.3 POWER SUPPLY AND I/O BOARD

3.4.3.1 NECESSITY

During the pre-study it was found that a constant 5V dc supply is needed as power supply to all the TTL voltage level circuits, and provision must be made for a constant 12V

supply which will be needed for the transmission link.

It was found that either 380V ac, 220V ac, or the 8,78V dc is available. This 8.78V dc is the voltage as taken from across the potential transformer side of the heater bridge inside the meter. This meant that some kind of conversion was necessary.

3.4.3.2 LAYOUT

- (a) Mode and meter selector
- (b) Mode- and disc rotation indicator
- (c) Disc revolution sensor
- (d) DC/DC convertor
- (e) Connectors between meter and interface
- (f) Connectors between interface and computer
- (g) Float preventor
- (h) AC/DC convertor
- (i) Phase 2 power down detector
- (j) Phase 3 power down detector

3.5 SOFTWARE

3.5.1 NECESSITY

The operation of the interface is dependent on controlling software. To cater for this very important aspect of the project, a software program was designed and developed in order to control the computer interface.

To be able to develop the controlling software, a study of the following aspects was required.

- (a) The operation of the existing thermal energy meters.
- (b) The way in which the different readings has to

be converted to digital data.

- (c) The way in which the digital data has to be controlled.
- (d) How to interface this data to a transmission medium.
- (e) The importance that time has to play during conversion and transfer of data.
- (f) The minimum facilities which the software has to provide.

3.5.2 Program layout

The essence of the program structure can be summarized under the following main headings:

- (a) Declarations
- (b) Interrupt
- (c) Initialization
- (d) Main program
- (e) Conversion routines
 - (1) Analogue-to-digital conversion
 - (2) BCD to ASCII conversion
 - (3) ASCII to hexadecimal conversion
 - (4) Hexadecimal to ascii conversion
- (f) Supply voltage level monitor routines
- (g) Reading generation routines
 - (1) Kilowatt hour reading generation
 - (2) Kilowatt hour set
 - (3) Maximum demand record clear
 - (5) Obtain readings
 - (6) Maximum demand calculation
 - (7) Arithmetic
- (h) Communication routines
 - (1) Data processing
 - (2) Data check
 - (3) Transmit record
 - (4) Message display
- (i) Synchronization
 - (1) Timing
 - (2) Led flash

- (j) Identification routines
 - (1) Cyclic redundancy check
 - (2) Meter identification

3.6 PROBLEMS EXPERIENCED

In order to improve the resolution of the circuit the 87C752 processor, which contains an onboard 8-bit A/D converter and 64 bytes RAM, had to be replaced with a processor which contains an onboard converter with a higher resolution and more RAM.

Because it is very difficult to calculate the maximum demand reading, more RAM was needed to improve the arithmetic calculation abilities of the processor in order to obtain more accurate readings.

3.7 FACILITIES

This project produced an interface between the mentioned meters and the computer via a select medium of transmission. It ensures that all the following readings can be obtained from the meter and displayed on a computer screen:

- (a) kilowatt hour reading
- (b) preset of kilowatt hour reading
- (c) instantaneous VOLTAMP reading
- (d) VOLTAMP reading
- (e) maximum demand reading.

3.7.1 FURTHER DEVELOPMENT

If additional features are needed in future, provision can be made for all the features provided by for example, the FULCRUM MULTIFUNCTION METER or a radio link.

In order to obtain the kilowatt

hour readings from dwellings, only a few software changes need to be made to service ± 16 meters from one computer interface.

A touch memory can be utilized to dump the necessary information from a group of meters into the computer. With the provision of an appropriate radio link, with a range of ± 4 km, all the necessary information could be dumped to different sub stations and transferred from there to a central point by means of a telephone line or radio link.

4.0 CONCLUSION

This is an excellent product if the price of $\pm$ R400 and the features of the interface and the conventional thermal Sangamo meters are compared to that of the modern electronic meter, which cost $\pm$ R3000. Not to mention anything about the fact that the existing, already proven reliable meters, could be successfully used with no modifications.

NON-INTRUSIVE MONITORING OF INDUSTRIAL ELECTRICAL LOADS

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ABSTRACT

A non-intrusive appliance load monitor determines the energy consumption of individual appliances turning on or off or operating under continuously varying load conditions. This monitoring system, which is implemented by means of neural network pattern identification technology, is based on detailed analysis of the current and voltage of the total load, as measured at the interface of the power source. The approach has been developed to simplify the collection of energy consumption data by utilities, but also has other applications. It is called non-intrusive to contrast it with previous techniques for gathering appliance data, which require placing sensors on individual appliances, and hence an intrusion onto the energy consumer's property.

1. INTRODUCTION

Researchers have long been interested in methods to enable them to monitor electrical circuits that contain a number of devices which switch on and off independently. Their interest rests primarily on gathering data as to the nature of the individual loads, the individual energy consumption, and other statistics such as the time of day an appliance in a circuit was on and off.

The primary application of gathered data is for load research purposes. Large electric utilities typically monitor dozens to hundreds of their residential customers with intrusive load monitors, placed on major loads such as electric heat, water heaters, refrigerators and air conditioning. The data is statistically averaged within demographic classes and used for a range of purposes by many audiences, including load forecasters, rate forecasters, public policy makers, and appliance designers [3].

A related application is the monitoring of individual utility customers for the purpose of an energy audit. These audits can be performed on the customer's request in order to analyse the characteristics of the appliances or the profile of daily/weekly/monthly energy consumption. The output of the audit would be useful in suggesting ways to reduce consumption and costs, which can be a considerable amount for large industry- and mining customers.

Other uses of power monitoring are for failure

analysis, security and operator-diagnostic purposes in industry and mining environments. By evaluating gathered data, ineffective operating practises can be identified and the necessary steps from management side can be taken to minimize or eliminate it.

Traditional load research instrumentation has involved intrusive techniques that require the installation of sensors on each of the individual components of the load and the use of complex data-gathering hardware but simple software. A monitoring point at each appliance of interest and wires (or sometimes power-line carrier techniques) connecting each to a central data-gathering location provide separate data paths, so the software merely has to tabulate the data arriving over these separate hardware channels.

During the recent years the emphasis in research has shifted to the investigation of non-intrusive methods to monitor electrical loads [3]. This approach reverses the balance to simple hardware but complex software for signal processing and analyzing purposes. With these methods no direct access to individual components of the load is necessary for installing sensors or making measurements. Only a single point in the circuit (eg. at the entrance of the main electrical feeder to an industrial plant) is instrumented, while mathematical algorithms must separate the measured load into separate components. In many load-monitoring applications this is a very cost-effective trade-off, which is a major advantage of this approach [3].

2. INDUSTRIAL ELECTRICAL LOADS

Various load models, that will be discussed briefly later, were developed in previous work [3,6]. These models concentrated on the on-off characteristics of appliances (residential) and did not cover appliances under continuously varying load conditions. These type of appliances are present in residential loads, eg. light dimmers, sewing machings and variable speed drills, which all have a continuous range of operating power levels and do not generate consistent on-off step changes in power consumption that could be fitted into the developed load models. As was outlined by Hart, this limitation to his models for residential load monitoring purposes was not significant because of the insignificant amount of energy consumed by this class of appliances.

However, these types of loads are much more likely to be present in industrial and mining environments, where e.g. variable-speed drives are more prevalent. Industrial loads can be categorized in two main levels, i.e. types of loads with alternating current, and types of loads with direct current.

- a) Loads with alternating current
 - i) An active load (mainly resistive) with a power factor, $\cos\phi > 0.996$.
 - ii) Loads with a lagging current, i.e. inductive loads, where $0^\circ < \phi \leq 90^\circ$, or $0 \leq \cos\phi < 1$.
 - iii) Loads with a leading current, i.e. capacitive loads, where $0^\circ > \phi \geq -90^\circ$, or $0 \leq \cos\phi < 1$.
 - iv) Distorting loads can be caused by non-linear elements in the circuit, e.g. by saturation phenomena in iron circuits of electrical motors and power transformers. Power electronic energy converters themselves also represent distorting loads. This type of load is characterized by the amount and magnitude of harmonic currents in the current waveform.
 - v) Owing to their starting and other dynamic characteristics, electrical machines represent a special type of load, the motor load.
- b) Loads with direct current
 - i) A resistive load corresponds to a load with an essentially resistive impedance R.
 - ii) An inductive load corresponds to a load with a L/R time constant ≥ 1.35 ms.
 - iii) A capacitive load corresponds to a load with a RC time constant ≥ 1.35 ms.
 - iv) In a load with counter-voltage, e.g. a battery load or a DC motor load, the AC content of the load current is much greater than the AC content in the load voltage.

During recent years power electronics have increasingly been applied in industry, mainly because of their technical properties and operating characteristics. Power electronic converters are employed in the entire field of electrical power engineering. The main applications are :

- i) industrial drives,
- ii) electrical power generation and distribution,
- iii) electrical heating,
- iv) electrochemistry, and
- v) electric traction.

Power electronic converters are employed in a power range extending from a few VA to over 1000 MVA. These static energy converters are characterized in accordance with their internal mode of operation, namely the type of commutation. This gives four main categories of energy converters :

- i) line-commutated converters,
- ii) load-commutated converters,
- iii) self-commutated converters, and
- iv) semiconductor switches and power controllers for AC.

Each of these categories will exhibit unique characteristics (signatures) when applied to fixed or continuously varying loads. Because these types of appliances contribute to a great extent to the total energy consumption in the industrial and mining sectors, it will be imperative to investigate their behaviour, characteristics and unique signatures in this proposed study.

3. APPLIANCE MODELS

In order to decompose the total load into its components, models of the individual appliances and their combination are needed. Electrically, the combination appears straightforward, as appliances are simply wired in parallel on a power system bus. Thus, to a first approximation, the power they consume is additive. This was the foundation on which Hart [3] based his research work. He proposed to examine the measured power vector (P & Q) of the load, and determine by heuristic procedures if a step change occurs.

Three classes of appliance models have been considered, in order of increasing generality (residential loads) :

- i) On/Off appliance model,
- ii) Finite State Machine (FSM) model, and
- iii) Continuously Variable model.

In the On/Off model an appliance is considered that may be either on or off at any given time, but allows for only a single type of ON state. This is a good model for most household appliances, such as a toaster, light bulb, or water pump. However, it makes no provision for electrically distinct types of ON states as found in a three-way lamp (low/medium/high) or washing machine (fill, agitate, spin). For this purpose a FSM model was developed, which allows for an arbitrary set of discrete states and state transitions, as found in heat pumps, dish washers and washing machines. Generalizing from the FSM model, continuously variable appliances have an infinite number of states. These types of appliances have actually a truly continuous range of operating power levels (like variable-speed AC drives) and do not generate consistent step-change signatures. Therefore they do not fit in the developed On/Off or FSM models. It is one of the author's aims to develop such a model for continuously variable industrial loads as it is not covered yet by other researchers in this field.

4. APPLIANCE SIGNATURES

Appliance signatures are the essence of the non-intrusive appliance load monitoring system [3,6]. Generally, an appliance signature can be defined as a measurable parameter of the total load that gives information about the nature or operating state of an individual appliance in the load.

A vector format is natural for representing signatures, in which each component can be a unique parameter of the appliance's signature. The two-dimensional area of appliance active power, P , on the horizontal axis, and appliance reactive power, Q , on the vertical axis, is used by both Hart and Sultanem as a "signature space" in which each appliance under observation will be located at a certain P-Q co-ordinate (two-component signature vector).

A non-intrusive signature is one which can be measured by passively observing the normal operation of a load, e.g. a step change in the measured power. Non-intrusive signatures are classified into two distinct classes, i.e. whether information about the appliance state change is continuously present in the load as it operates ("steady-state signatures") or only briefly present during times of state transition. The power, current or admittance of the total load at the utility fundamental frequency can be measured and step changes can be identified as signatures. Additional information can be obtained by examining the harmonic currents generated by appliances. The main contributor to the generation of harmonic currents in the industry is electronic power converters. Each category of power electronic converters (e.g. line-commutated converters) generates harmonic currents of some order and magnitude, and this information can be used as signatures to identify the appliances.

Under non-linear load conditions it is impossible to describe power only by two orthogonal components, which is generally known as the active power component, P , and reactive power component, Q . A third component, called distortion power, D , is introduced so that electric power consists of three mutually orthogonal components, P , Q and D . Under steady-state conditions the three-component (P, Q, D) signature vector of a certain appliance will be located somewhere in the P-Q-D power-space.

Transients in consumer appliances appear to come in different shapes, corresponding to the generating mechanism. A typical example of an appliance that exhibits a unique switching-on transient, is a three-phase induction machine. Parameters for categorizing transients are their size, duration, time constants, or parametric variables in models which can be fit to the observed waveforms. However, other researchers have not pursued transient signatures as signatures in their prototype monitoring equipment, but relied on steady-state signatures. One reason given is that transients are variable in parameters like duration, shape and size, because it often depends on the exact point in the voltage cycle at which the turn-on or turn-off occurs.

5. PRACTICAL IMPLEMENTATION AND RESULTS

The problem of appliance signature identification can be solved by using trained artificial neural networks. The majority of neural network applications use three layer networks with learning by Back-propagation [1]. This multi-layer perceptron can be trained as desired by using analytical functions for the activation of the network nodes ("neurons") and by applying a backward error propagation algorithm to update the interconnecting weights and thresholds until proper recognition capability had been attained [2,5]. The components of the N-component signature vector can be obtained by processing of the measured load current and voltage waveforms, from where unique appliance signatures can be identified (character extraction), e.g. steady-state signatures like orthogonal power components P , Q and D , harmonic current components (order and magnitude), transient characteristics, time of use, shapes of P-Q-D power space trajectories under varying load conditions, etc.

For research purposes a three phase measuring instrument was developed to measure the three phase load currents and voltages at a single point. This instrument consists of three 300 Ampere "LEM" current modules and three 380 Volt "LEM" voltage modules (Hall effect sensors) and provides isolated $\pm 5V$ analog signals to a PC-30D analog I/O card that slots into an IBM PC. The voltage and current signals are processed and the instantaneous values of active (p) and reactive (q) power [7] are calculated in real time using :

$$\begin{aligned} p(t) &= v_a i_a + v_b i_b + v_c i_c \\ q(t) &= -[(v_a - v_b)i_c + (v_b - v_c)i_a + (v_c - v_a)i_b] \frac{1}{\sqrt{3}} \end{aligned} \quad (1)$$

In the presence of non-linear loads these two calculated powers can each be separated into a mean value ("—") and alternating value ("~") part as follows:

$$p = \bar{p} + \tilde{p} \quad q = \bar{q} + \tilde{q} \quad (2)$$

Watanabe [7] shown that $\bar{p} = P$ and $\bar{q} = Q$, where P and Q are the conventional three phase active and reactive powers respectively. He further defined the average deactive power D as follows:

$$D = \sqrt{\bar{p}^2 + \bar{q}^2} \quad (3)$$

where $\bar{P}$ and $\bar{Q}$ are the RMS values of $\tilde{p}$ and $\tilde{q}$ respectively. The result of these powers can be written in the well known form as :

$$S = \sqrt{P^2 + Q^2 + D^2} \quad (4)$$

and is normally represented using the power tetrahedron, shown in figure 1, instead of the triangle as in the linear case.

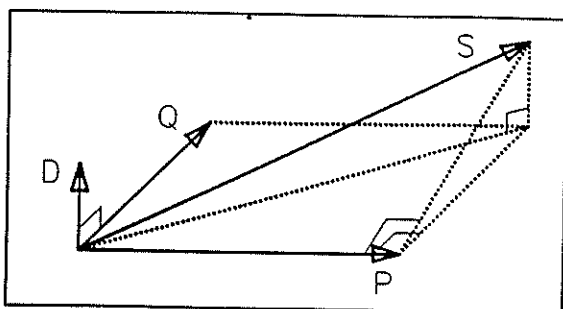


Figure 1 : Power Tetrahedron

With the aid of the developed hardware, trajectories of the three orthogonal power components P, Q and D were obtained for both a 30 kVA induction motor and a 100 kW DC motor driven from a 6-pulse AC/DC converter. The load of both appliances were varied from 0 - 100%, and the results obtained are given in the form Q & D vs. P in figures 2 and 3 for the induction motor and DC drive respectively.

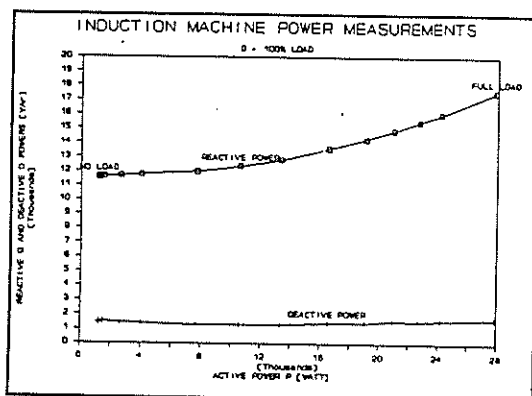


Figure 2 : Induction Motor Power Curves

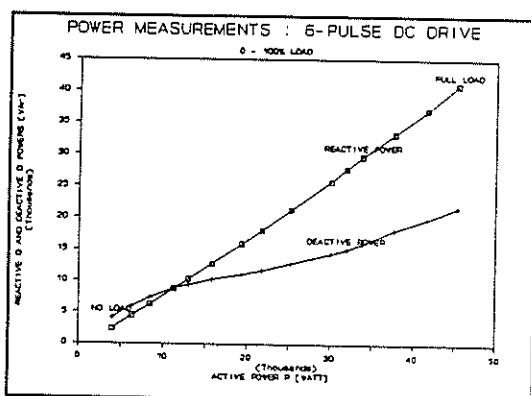


Figure 3 : DC Drive Power Curves

These are the latest results obtained by the author. It is evident from these results that each appliance

exhibits unique power curves over the full loading range. Future work includes the determination of similar power curves for other typical industrial appliances under varying load conditions. From these results unique characteristics have to be identified (character extraction), e.g. the slopes of the curves $\Delta P, \Delta Q, \Delta D$ etc. Over the full loading range these curves ($P, Q, D, \Delta P, \Delta Q, \Delta D$) will be divided in a number of steps (say e.g. 20) to form 20 sets of input vectors to the neural network pattern classifier. These sets will be used as input vectors to train the neural network.

6. SUMMARY

A non-intrusive appliance load monitor is proposed in this paper to determine the energy consumption of individual industrial electrical appliances under on/off or continuously varying load conditions. This identification is based on detailed analysis of the current and voltage waveforms of the total load, as measured at the interface to the power source. Neural networks will be used as pattern classifiers to identify the various appliances.

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LOAD RESEARCH STRATEGY AND DEMAND SIDE PLANNING MODEL FOR DOMESTIC WATER HEATING

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Abstract

The problem of obtaining reliable data concerning the different end use categories in the domestic sector is addressed in this paper. A load audit and data processing strategy that enabled the authors to produce a disaggregated load profile for the domestic sector is discussed. This load profile included a load curve for the geyser load. A modelling technique was applied to this load profile to determine what the impact of instant water heaters, different thermostat setting and element sizes would have on it.

Introduction

There is a growing need to gain information on the demand of electricity by the domestic sector. Such information may be useful for planning of the generation and distribution of power and to point out demand side management potential in the different appliance categories.

For these reasons a project was initiated to obtain accurate data on the different categories of end uses that contribute to the domestic load profile. This data collection process had to be done in a cost effective way.

In order to determine the usefulness of the data that was obtained, modelling techniques were applied to one of the categories namely water heating. Using this data, simulations were done to determine the effect of the following on the load profile:

- 1) Instant water heaters.
- 2) Thermostat setting.
- 3) Element size.

Measuring methodologies

In order to obtain data of the kind that was mentioned before one would have to measure on a household level. The options available to the load auditor make use of centralised and / or decentralised measurements. The nature of these options is discussed in the following paragraphs.

Centralised measurements

If different end use categories have to be measured in a house, the obvious place to do this is at the distribution board (DB). The different categories that can be measured at this point are the stove, lights, plugs, geysers and pumps (when present). A multi-channel data logger is suitable for collecting time related energy data at the DB.

Decentralized measurements

Decentralized measurements imply measuring at different locations at appliance level. This will allow one to divide the plug category into smaller relevant categories like: Space heating, cool storage, lighting, food preparation and life quality appliances.

Equipment suitable for recording such data is less readily available. One such system on the market is the "Collective Appliance Monitoring System". This system consists of a load sensing element which is introduced between the plug of the appliance and the wall socket, as well as a data collection element. The data collection element receives measurement data from up to sixteen load sensing modules via radio transmission.

Data collection method used

The aim of the data collection process was to disaggregate the domestic load profile in as many useful categories as possible. In order to keep within budget limits, it was decided to do only centralised data recording. An expansion of the appliance categories with the use of load audits done by means of questionnaires was considered. A discussion of the audit process is however not in the scope of this paper.

The data collection process

An area in Lynnwood, Pretoria, fed from one transformer was targeted for load audits. Eight houses (from a total of 39) in this area were selected for the installation of data loggers.

Nature of the Data Loggers

The data loggers which were used have sixteen analog input channels with a range of 0-200 mV and register

time averaged values for selectable intervals of between one minute and one hour. Clip-on CT's with the desired output range were used in conjunction with the loggers to register currents.

Deciding on an averaging time interval

To improve the information quality of the data as far as possible, a short time interval is required. There are limitations on the memory of the data loggers and to avoid bothering the inhabitants too often, a compromise had to be found. For a five minute interval it was only necessary to visit the equipment once a month and this was considered acceptable.

Processing of Data

After three months of collecting data, well over a hundred thousand data samples were collected from which a disaggregated load profile had to be constructed. Information on how to analyze this type of data could not be found in the literature. This is mainly because researchers previously working in this field, for instance [1], made use of hourly data of total household loads. In [2] the use of separately measured data is mentioned but it is not stated exactly how this data was analyzed.

Separating the load

To separate the load profile for each house was not as simple a task as one would expect. The reason for this is that often the geyser, or some times geysers, as well as the swimming pool pump had to be measured together with other loads.

Another factor that complicates matters is the fact that most houses had some kind of sub distribution board. These sub DB's often feed distant parts of the house or living spaces separated from the house. For this reason when possible a channel was devoted to measure the current that fed distant DB's.

A load separation algorithm was developed to separate geyser loads from combined data [3]. The result of one such algorithm that was used to separate the geyser load from recordings of a sub DB is shown in Fig.1. Unfortunately the loads of only five of the houses could be separated completely.

Construction of the Disaggregated Load Profile

The construction of the disaggregated load profile was finally done by combining the separated loads of the five houses.

Accounting for Diversity

Diversity of different households is a well known topic and the way to handle this was simply by averaging

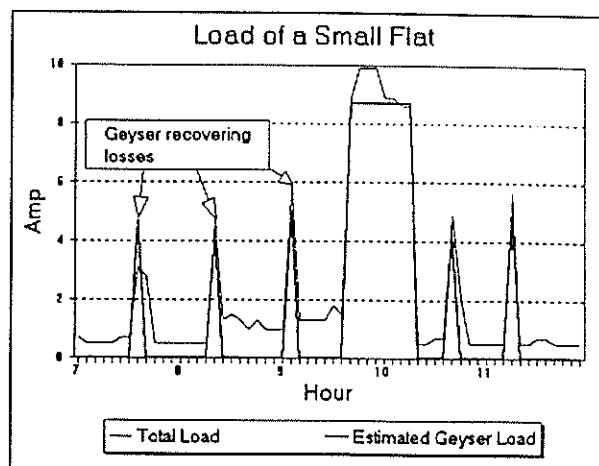


Figure 1: The result of the algorithm which identifies the geyser load from a combined load profile.

data recorded at the same time and date for different houses.

Averaging of all Winter Days

Although it can be expected that for one house the load, for example for a Monday, would have the same pattern for most Mondays, surely they would not be identical. It was found that these differences could be evened out by averaging the load after diversity for every individual day of the week.

Results

After each of the end use categories were put through the process that was described above, the results indicated in Fig.2 were obtained.

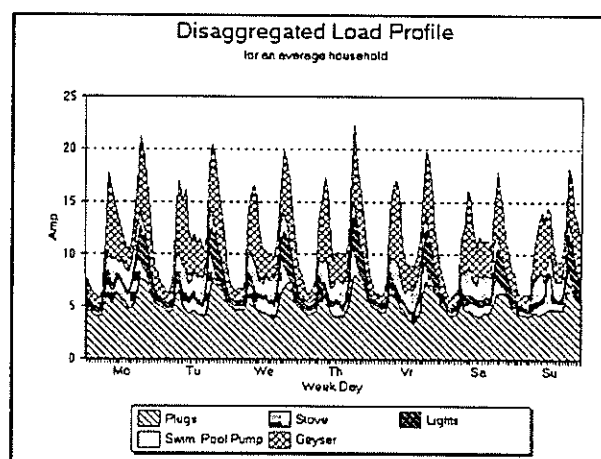


Figure 2: Disaggregated Load Profile

Evaluating the Load Profile

The estimation process was evaluated by comparing the transformer load of the transformer that feeds the area's 39 houses with the estimated load, after it was multiplied with a factor to reflect the same energy use.

This is shown in Fig.3. The correlation is quite remarkable considering the small sample of 5 out of 39 houses.

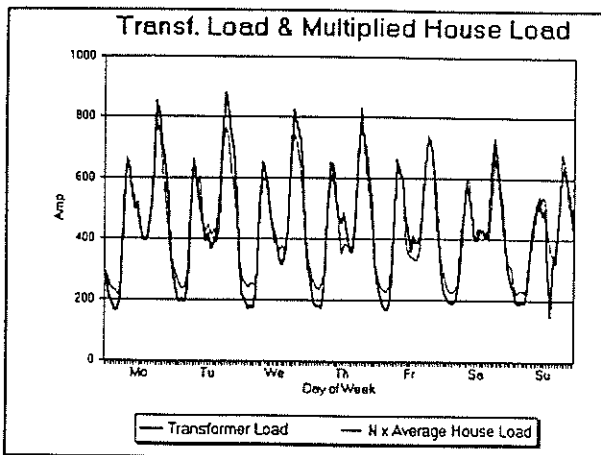


Fig. 3: Comparison between the estimated transformer load and the actual load.

Demand side Planning Model for Domestic Water heating

The main influence on the energy usage pattern of domestic water heaters are the following: Water inlet temperature, thermostat setting, water usage pattern, water heater element size and the type of water heater.

It was attempted to determine the effect of a variation of these parameters on the energy usage pattern. For this purpose a model which was developed by Beute [4] was adopted. This model is based on the following equation:

$$P(t) = N_{TOT} \cdot P_e \cdot (F(t) - F(t - t_{on})) \quad (1)$$

where: $P(t)$ = Energy used at time t in Watt.

$F(t)$ = The sum of one or more probability distribution functions that estimate the load profile.

t_{on} = The time a geyser element is switched on. This depends on the average element size, thermostat setting, inlet temperature and amount of water extracted.

P_e = The average element size.

N_{TOT} = The number of geysers.

For more information on the theory of this model please refer to [4].

Implementation of the Model

A Computer Program was written to determine the parameters of $P(t)$ that will optimally simulate the measured load curve.

$F(t)$ was constructed by combining three functions: a constant function that represents the standing losses in

geysers, and two Weibull functions that simulate the morning and evening loads. The Weibull function was selected because of its ability to take on an unsymmetrical appearance.

Results of Simulation

In Fig.4 the actual and simulated load curves are compared. The following parameters were assumed:

Average element size = 3 kW

Average Thermostat setting = 65 °C

Average Inlet Temperature = 20 °C

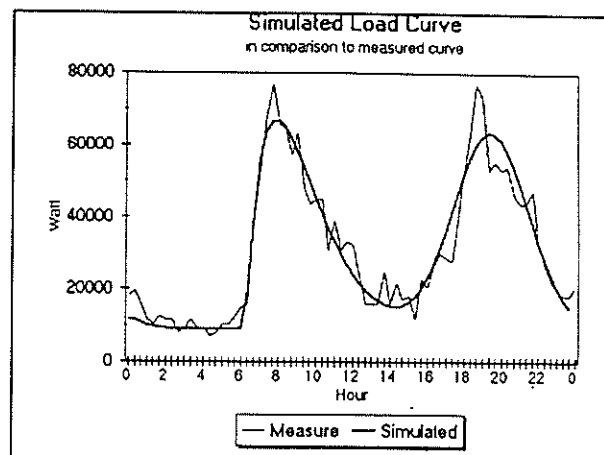


Fig. 4: Comparison between actual and simulated curves.

Fig.5 shows the three different components of $F(t)$.

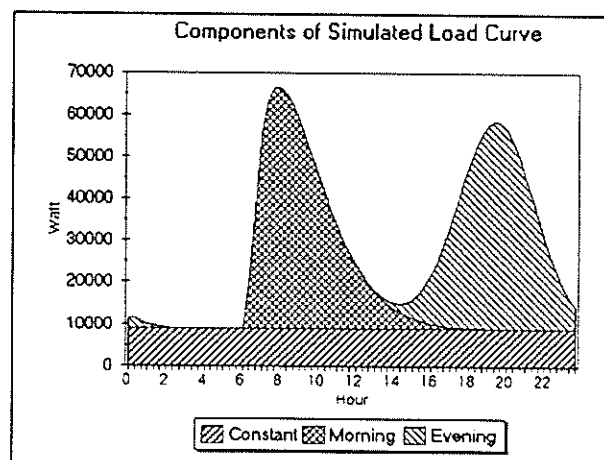


Fig. 5: Different components of simulated curve.

Size of Geyser Element

If the size of the element is changed to 1kW, the geyser will take longer to heat the water up to thermostat setting. The peak loads is not that high though. This can be seen in Fig. 6.

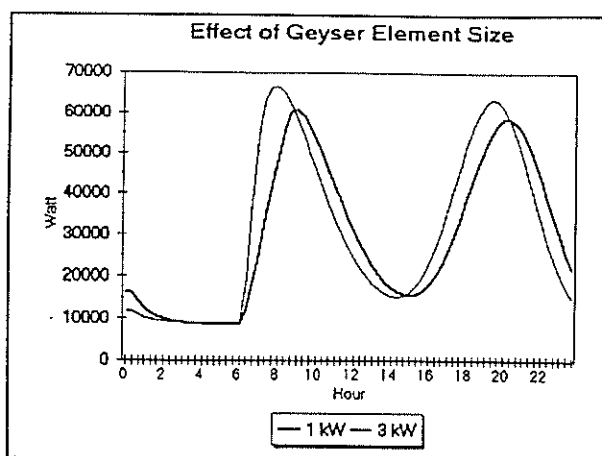


Fig. 6: Change of element size.

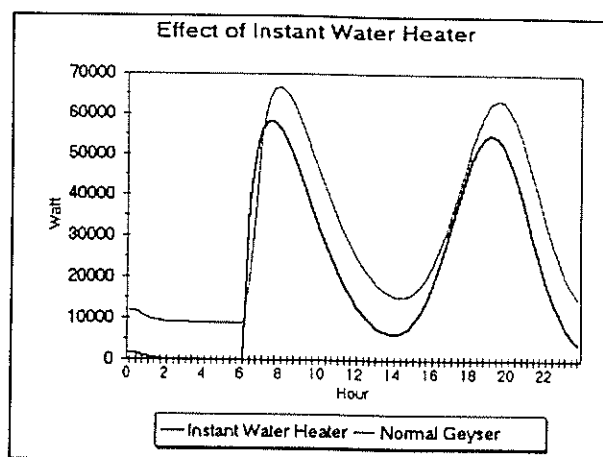


Fig. 8: Instant Water heaters.

Thermostat Setting

The effect of changes in thermostat setting can be seen in Fig 7.

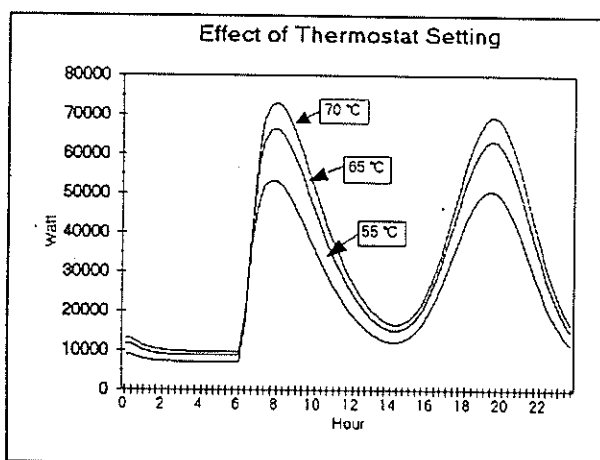


Fig. 7: Change of Thermostat setting.

Instant Water Heaters

If instant water heaters are simulated, the constant function in the simulated data must be set to zero, because there are no standing losses when using instant water heaters. Further, the element size must be set to 30kW. The result can be seen in Fig. 8.

Conclusion

Fig. 3 leads us to the conclusion that the load audit strategy that was used in this project proved to be a cost effective way to obtain a representative disaggregated load curve for an homogeneous residential area.

The data obtained can be useful for modelling exercises and was applied successfully to a load model capable of simulating the effect of some variable parameters on the water heating load profile.

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ACKNOWLEDGEMENT

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Evaluation of domestic solar water heating systems

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Abstract

It is a well known fact that the energy consumption for the water heating of an average household can be up to 40% of the total energy consumption of that house. The need to minimise this high consumption is growing. Solar water heating (SWH) systems offer a solution to this problem, although the current information about SWH systems is relatively limited. If more concrete proof could be found to convince the consumer, these appliances would definitely gain popularity.

Introduction

The purpose of this study can be summarised in two main objectives, namely

- i) The accurate determination of the effect of a domestic solar water heating system on the energy consumption of a house; and
- ii) The providing of an economic evaluation method or life cycle cost (LCC) model for SWH systems.

The final goal is to evaluate, with the help of the LCC model, different types of SWH systems in terms of a specific hot water usage pattern. In other words, a specific consumer's hot water usage is simulated on a SWH system. From the results obtained from this simulation it will then be possible to determine whether or not that specific SWH will be profitable after a number of years.

The effect of a SWH system on the energy consumption of a geyser.

Previous LCC models that have been developed for SWH systems, depended on approximations of losses and energy savings of these systems. It is difficult to prove that SWH systems are necessarily an economic proposition. This is primarily because of the fact that SWH system efficiency is a function of the consumer pattern. In other words, no average effectiveness can be determined, because each consumer will show a unique saving or loss.

With this research an easy solution to overcome this imperfection in the most cost-efficient manner was sought. A way had to be found to determine exactly

how much a consumer would save with a SWH system without physically installing a SWH system.

In an attempt to solve this problem, the following strategy was followed:

instead of taking the SWH system to the consumer, the consumer's hot water usage pattern was taken to the SWH system. In other words, the consumer would now be able to determine exactly how much he might save on any type of system without installing the system. A way had to be found to simulate different types of hot water usage patterns on SWH systems. The electric energy usage of the geyser element had to be accurately logged.

It had been decided to do the simulation with an electrically controlled valve and Pulse Width Modulator (PWM). The reason for the PWM will be discussed later. The control would be done with a personal computer supplied with an A/D and D/A converter.

To log the energy consumption, there had been decided to make use of the SAMES SA9102C chip which measures single phase power/energy and gives a frequency output in the form of pulses that are directly proportional to the energy used.

A preliminary design of a part of the system is shown in Fig. 1.

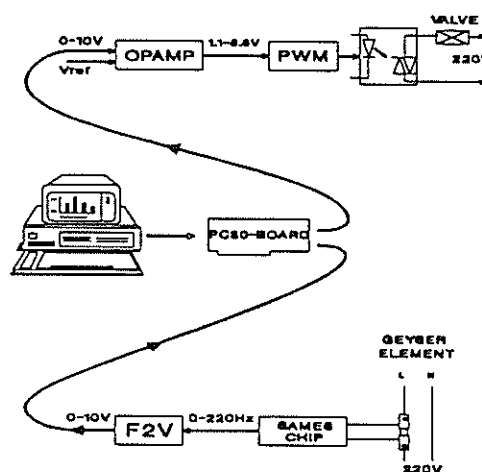


Figure 1: General design

A diagram is shown in Fig. 2 that illustrates how the

total system is currently connected for the simulation and data logging functions.

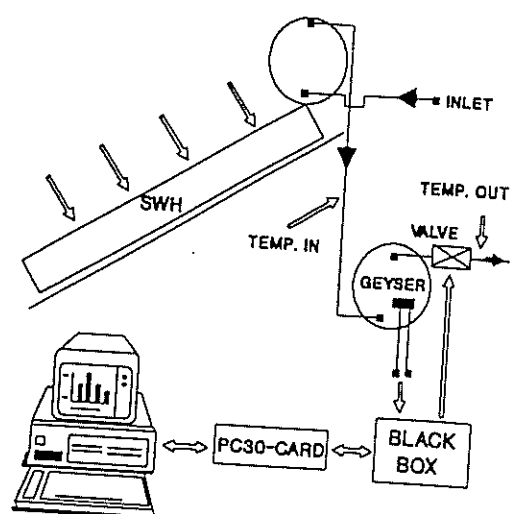


Figure 2: Experimental configuration

The PWM was mainly for research reasons. This research is aimed at determining what the effect on the electric energy usage of, for example, 25 000 houses would be if all of them were supplied with SWH systems.

Through calculation the usage pattern of 25 000 houses can be adapted to the usage of only one geyser. Because many houses use hot water simultaneously, this usage is continuous and not discrete like the single house case is. The PWM is then used to simulate this usage on the SWH system.

A Turbo Pascal programme has been developed to control the in- and outputs to and from the PC-30 card. Before the programme starts, a data-file has to be generated for programming the hot water usage that has to be simulated. This data is divided into five minute intervals and thus occupies 288 rows. By editing this data-file, different types of usage patterns can be simulated.

In an attempt to experiment with as many SWH as possible, the Department of Agriculture and Agricultural Economy gave their consent for the experiments to be conducted at their research centre, Silverton, Pretoria. Currently there are approximately 10 different types of SWH systems.

Results

Because of a time limit, the research was preliminary limited to only one type of SWH. The results obtained up to this stage are extremely interesting and

successful. In Fig. 3 the energy usage of the geyser element without a SWH system with the consumer pattern, as shown in Table 1, is shown. Fig. 4 shows the energy usage of the same geyser element, this time with a SWH system.

Table 1

Time	Litres used	Usage
6:30	38	
6:45	2	Shower
6:50	2	Sink
6:55	4	Sink
8:00	8	Sink
9:30	76	Sink
12:30	2	Laundry
17:30	4	Sink
18:30	2	Sink
18:45	6	Sink
19:00	15	Sink
19:05	15	Dishwasher
19:10	15	Dishwasher
21:30	57	Dishwasher
22:00	4	Bath
22:30	2	Sink
		Sink
Total	250 Litres/day	

Energy usage of geyser without a SWH system

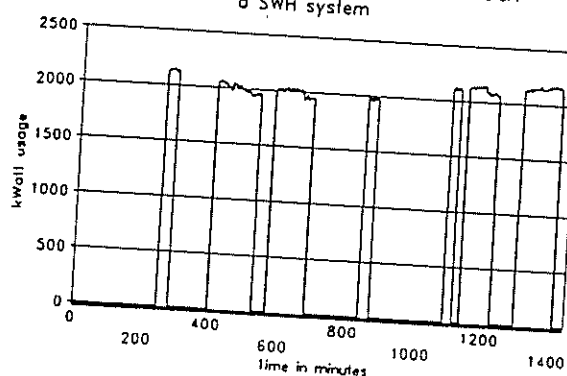


Figure 3: Energy usage of geyser without a SWH system

Energy usage of geyser with a SWH system

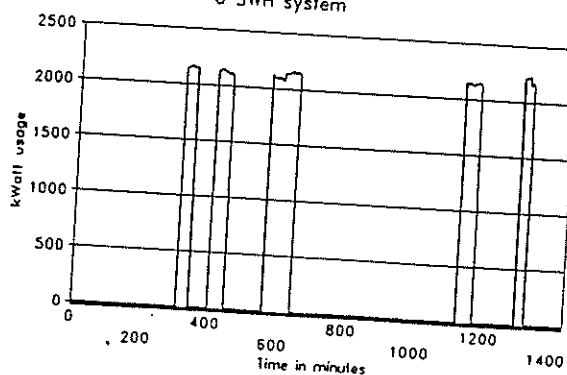


Figure 4: Energy usage of geyser with SWH system.

The in- and outlet water temperatures are shown in Fig. 5 and 6 respectively.

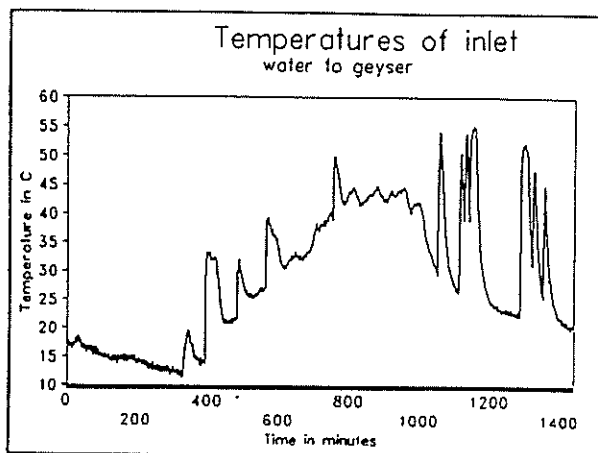


Figure 5: Inlet water temperature

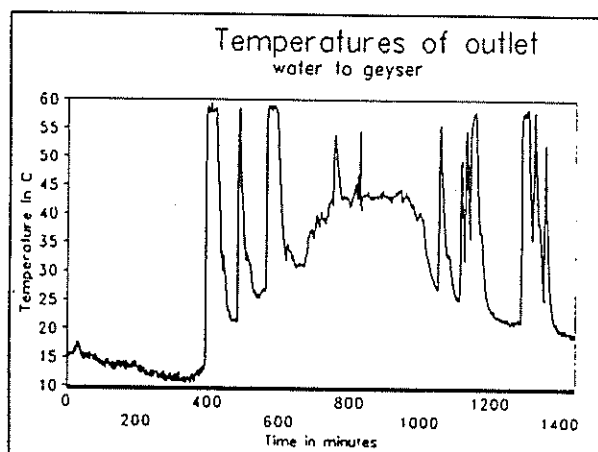


Figure 6: Outlet water temperature

The reason for the sudden increase and decrease in temperatures if water is released, is because of the fact that the temperature sensors are situated in the pipes and not in the geyser.

By calculation from data shown in Fig. 3 and 4, the percentage energy saved by the SWH can be determined. This saving is dependent on the weather climate and type of consumer. For the specific case as in Fig. 3 and 4 the saving was approximately 52.6%.

Economic evaluation of SWH systems

It may be debated whether SWH systems are necessarily economically viable. Since many householders have no idea of how much they should invest in a SWH system, an economic evaluation procedure or an indication of a reasonable price is required.

In practice, the likely cost savings are difficult to estimate because they involve predictions of what will happen in the future. One argument based on "The Time Value of Money", explains this in the following way:

"If I invest a sum of money now in a solar water heater because of the reductions this will bring about in my energy bills, I want to be reasonably sure that I could do no better by, for example, investing the same sum of money in a building society and using the interest to pay the increased electricity bills. Inflation reduces the value of money invested in this way and must be taken into account."

Further factors that have to be taken in account for the overall evaluation are:

- the time value of money which is removed by discounting future sums of money to the present by means of an appropriate discount factor, often the long term interest rate.
- the increase in energy costs
- the economic life of the SWH system that can be anything between 10 and 20 years
- the tax and fiscal incentives
- the incremental cost of SWH systems and
- the percentage saving as determined by the simulation.

The calculation technique which attempts to take these and other factors into account is called "Life Cycle Cost". The technique is graphically illustrated in Fig. 7. [1]

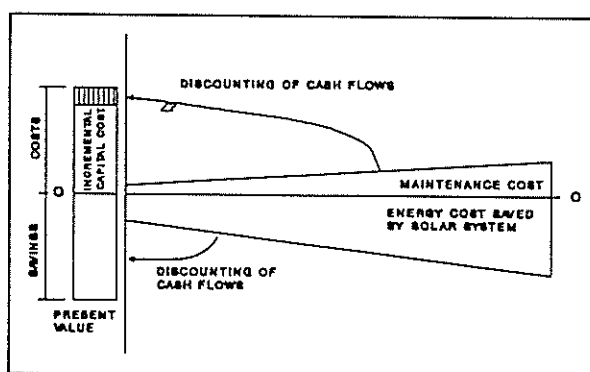


Figure 7: Life cycle cost evaluation of SWH systems.

With the above mentioned method it was possible to determine a LCC model for the system currently tested.

The following figures were taken in account:

Long term interest rate : 11%

Inflation rate : 10.8%
Energy cost : 15.59 c/KWh
SWH cost : R5900
Economic life of SWH : 15 years
Percentage saving : 52.6%

From this and other factors that have been taken in account it could be determined that the total income for the current SWH would be approximately R560 more than that of an long term investment.

Conclusion

A SWH system can not be judged on the results of two or three tests. In order to make the most accurate predictions it is a necessity to get as many test results as possible of a specific SWH system, simulated in as many different conditions as possible.

Solar water heaters are available which can provide hot water for domestic uses effectively and reliably. In most areas in South Africa the installation of solar water heaters is an economic long term proposition.

Once home owners can be convinced that solar water heaters really are a good investment, these devices will gain popularity. In this way, a contribution can be made to the national economy by reducing the use of our limited fossil energy resources and/or by decreasing the national peak electric load.

References

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A MINIMUM DISCOMFORT-MINIMUM ENERGY CONTROLLER MODEL FOR DOMESTIC WATER HEATING

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ABSTRACT

A two-dimensional model for the domestic water heater has been developed by Rautenbach and Lane[1]. This model will be used to develop and test the controller model. The performance of this water heater model is discussed. The structure of the controller model is presented in the second part of the paper. Then follows a discussion of the performance of the controller model, as well as future work to be done.

1. INTRODUCTION

The controllability of domestic appliances has become an important field of study in recent years. Because of this, the development of models describing these appliances has become necessary to test the performance of various control algorithms. These control algorithms have the following in common: to minimize cost for the utility and customer, without causing discomfort for the customer. The discomfort for the customer can be quantified as the lack of hot water when demanded. By reducing the peak electricity demand, the utility may save the cost of building a new power station. If a dynamic pricing structure exists, this may also cause savings on the part of the customer.

Controlling options for the domestic sector are currently limited, mainly because of the existing flat rate pricing structure. There are no incentives for the customer to reduce electricity usage during peak times, resulting in high peaks with increasing costs for the utility.

Therefore, the development of the controller model is aimed at a dynamic pricing structure, such as a time-of-use tariff structure or spot pricing. This allows possible cost savings for all the parties involved.

2. WATER HEATER MODEL

The model is based on the physical data attained from monitoring a domestic water heater. In order to keep the controller as simple as possible, the authors decided on a two-dimensional model for the water

heater (or 'plant' in control terminology), with the inlet and outlet temperatures the states of the plant.

2.1 LINEAR DIFFERENTIAL MODEL

The linear differential model is then described by the following equation:

$$\dot{T}(t) = \begin{matrix} A(t)T(t) + B(t)F(t) + \\ C(t)U(t) + D(t) \end{matrix} \quad (1)$$

where:

$T(t)$ = States (temperatures) of the plant (water heater).

$F(t)$ = % Flow rate of the incoming water (e.g. 0, 50, 100).

$U(t)$ = One when the element is on, zero otherwise.

$A(t)$, $B(t)$, $C(t)$ and $D(t)$ are time dependent state matrices, containing the coefficients of the linear differential equations, which change for each of the following operating conditions:

- 1) $F(t) = 0$ and $U(t) = 0$
- 2) $F(t) = 0$ and $U(t) = 1$
- 3) $F(t) = 50\%$ and $U(t) = 0$ or 1
- 4) $F(t) = 100\%$ and $U(t) = 0$ or 1

A graphical representation of the temperature profiles for conditions 1 and 2 are shown in Figure 1 and for condition 4 in Figure 2. The inlet and outlet temperatures are shown in both cases.

A curve fitting procedure was performed on the data obtained from the water heater. It is apparent that only two types of curves are needed to describe the temperature profiles: a first order exponential decay and a straight line, respectively, of the form:

$$T_1(t) = \alpha e^{-\beta t} \quad (2)$$

$$T_2(t) = At + B \quad (3)$$

Equations 2 and 3 translate to the differential domain in the following manner:

$$\dot{T}_1(t) = \beta T_1(t), \quad T_1(0) = \alpha \quad (4)$$

$$\dot{T}_2(t) = A, \quad T_2(0) = B \quad (5)$$

Equations 4 and 5 yield the coefficients used in Equation 1 to describe the temperature profiles inside the water heater.

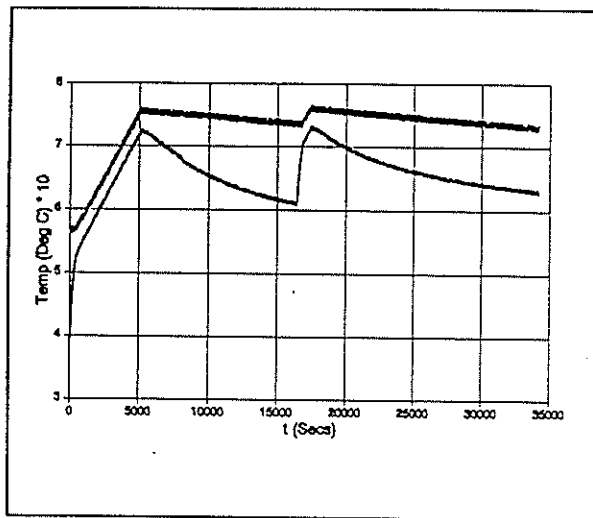


Figure 1: Data of the water heater (tap closed and element switching).

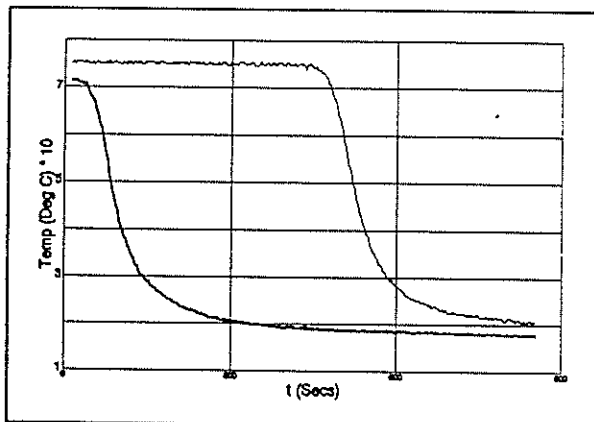


Figure 2: Data of the water heater (tap open 100%).

In the simulation model of the water heater, operating conditions 2,3 and 4 are divided into sub-conditions, to account for the change in dynamics of the temperature curves within these conditions. The switching of coefficients in condition 2 is controlled by the difference between the inlet and outlet water temperatures. In conditions 3 and 4 the amount of energy in the tank controls the switching between

coefficients. Energy drawn from the water heater can be calculated as follows:

$$P_{out} = \dot{m}_l C_p T \quad (6)$$

where:

- P_{out} = Rate of energy removal from the water heater [J/s].
- $\dot{m}_l$ = Rate of water flowing out [kg/s].
- C_p = Specific heat capacity of water [J/kg-°C].
- T = Temperature of water flowing out [°C].

The rate of energy returned to the water heater is equal to the element rating, in Joules per second.

Figures 3 and 4 show the output of the model for conditions corresponding to those in Figures 1 and 2. These results are clearly very satisfactory.

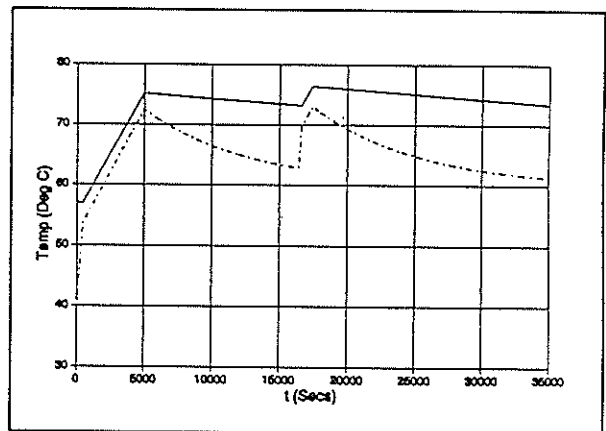


Figure 3: Simulation of the water heater (tap closed, element switching).

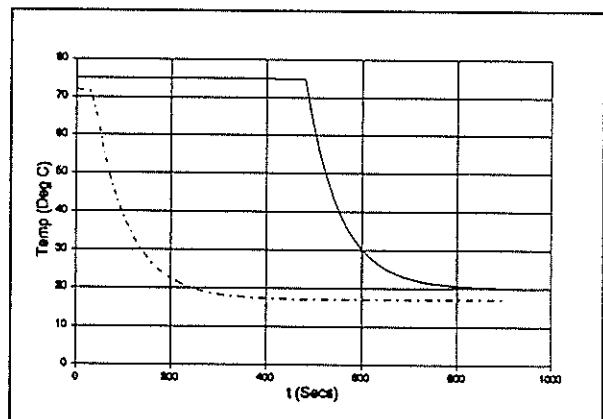


Figure 4: Simulation of the water heater (tap open 100%).

It is necessary, however, to note that the coefficients of the differential equations (and subsequently the

graphs) would change when the environment of the water heater changes (e.g. the inlet water temperature drops). The influence of these changes is predictable and can be accommodated in one of the following ways: compile a database to cover most cases; or describe each coefficient as a function of the various influencing factors (e.g. a function of inlet and ambient temperature, etc.).

The second option is much more elegant and is currently being investigated by the authors.

3. CONTROLLER MODEL

As mentioned previously, the controller is developed with the aim to minimize cost and discomfort. This is possible when a dynamic pricing structure, such as the time-of-use tariffs depicted in Figure 5, exists.

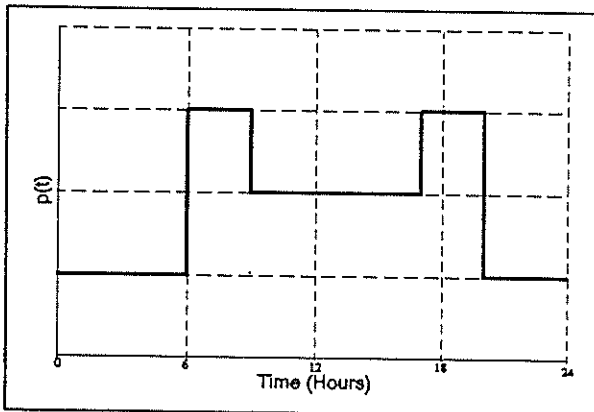


Figure 5: A time-of-use tariff structure for one day.

The cost minimization is achieved by reducing the usage of electricity when it is more expensive. The optimal case would be not to use electricity at all during these times, but this would obviously effect the comfort of the customer negatively. By storing energy in the water heater during times of less expensive electricity, the minimization of cost can be achieved, while causing the minimum discomfort to the customer.

Hence, two separate cost functions are minimized by the controller- one when electricity is cheap, the other when it is not.

3.1. COST FUNCTIONS

a) Inexpensive electricity periods

The energy cost for the time interval $[0, t_f]$ can be defined as

$$C = \int_0^{t_f} Q_{IN}(t) dt \quad (7)$$

where $Q_{IN}(t)$ is equal to the energy added to the heater at time t .

The optimal control problem can now be formulated as follows: Find the set of optimal element switching times so as to minimize the cost function C , subject to the following temperature constraint

$$T_{OUT}(t) > T_{MIN}(t)$$

and the final time constraint

$$T_{OUT}(t_f) > T_{FIN}(t_f)$$

where:

- $T_{OUT}(t)$ = Outlet temperature.
- $T_{MIN}(t)$ = Desired minimum outlet temperature.
- $T_{FIN}(t)$ = Desired minimum outlet temperature at the final time.

The second constraint is responsible for the energy storage. Using penalty function formulation, a new cost function can be defined as follows:

$$J_1 = c_1 \int_0^{t_f} Q_{IN}(t) dt + c_2 \int_0^{t_f} G(t) dt + c_3 H(t_f) \quad (8)$$

where $G(t)$ is a function which is equal to zero when the temperature constraint is met, and proportional to the difference between $T_{OUT}(t)$ and $T_{MIN}(t)$ otherwise. Similarly, $H(t_f)$ is equal to zero when the final time constraint is met, and proportional to the difference between $T_{OUT}(t_f)$ and $T_{FIN}(t_f)$ otherwise.

b) Expensive electricity periods

When the price of electricity rises, the cost function alters slightly. $H(t_f)$ is removed and the cost function becomes:

$$J_2 = c_1 \int_0^{t_f} Q_1(t) dt + c_2 \int_0^{t_f} G(t) dt \quad (9)$$

By increasing c_1 in Equation 9 the electricity usage during these periods will be minimized. The values of c_1 and c_2 in Equation 8 and 9 can also be varied so as

to differentiate between income classes. For a higher income group c_2 might be increased with respect to c_1 so as to accentuate discomfort minimization. For a lower income group the opposite is true.

An initial set of element switching times has to be specified. These switching times are then shifted around by the controller, resulting in a minimization of the cost function.

3.2 EXPERIMENTAL RESULTS

The parameter optimization technique described by Snyman [2] was used to optimize these cost functions. The result is an open loop control for the element in the water heater. Some results are shown in Figures 6 and 7.

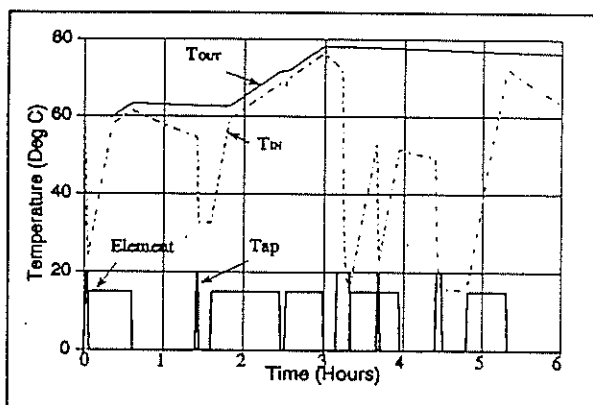


Figure 6: Output of the controller model (five element switching periods).

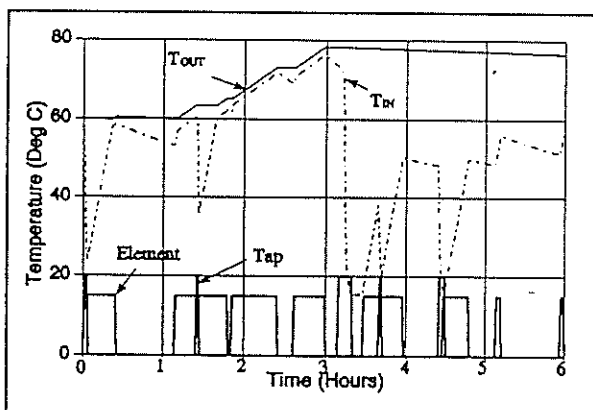


Figure 7: Output of the controller model (eight element switching periods)

For both simulations the electricity was assumed to be less expensive during the first three hours and more expensive the rest of the time. In the first simulation (Figures 6) five initial element switching times were assumed and in the second (Figure 7) eight. The same usage pattern was used for both cases.

The final time temperature constraint for the inexpensive period cost function (after three hours) was set at 75 degrees C, with the running time temperature constraint (T_{MIN}) set at 65 degrees C for the outlet water temperature.

The potential of this type of open loop control can be seen in the fact that energy is stored when electricity is inexpensive, reducing electricity usage when the price of electricity rises (peak load hours). The possible cost savings can be determined by calculating the amount of electricity used and comparing this with usage under flat rate tariffs. A marked difference in the electricity bill should also appear.

To find the optimal element switching pattern a series of simulations with different usage patterns have to be run, so as to arrive at an 'average' switching pattern. This process is currently under investigation. If the results are positive, it is hoped that electricity distributors will find it beneficial to implement a more suitable control strategy under dynamic pricing conditions.

4. CONCLUSION

In this paper a controller model for the domestic water heater was presented. The aim of this controller is to decrease cost for the customer and utility, while keeping the discomfort for the customer at a minimum. Some preliminary results showed the potential of this controller model. A reasonable amount of research still has to be done, but it is hoped that the model can be implemented in the near future.

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COST-EFFECTIVE DATA COLLECTION AND CONTROL OF DOMESTIC ELECTRICAL ENERGY USAGE.

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ABSTRACT

The current and proposed electricity tariff structures have shown a definite trend towards time-of-usage based tariffs. The key element in such time-of-use tariffs, is real-time information. We require real-time information regarding the energy consumption of a given user, at a particular time. This information is easy to obtain, but the difficulty lies in the gathering of this information. Power Line Carrier (PLC) although not a new concept, requires a cost-effective implementation to allow its use in a real-time data gathering system. The proposed system makes use of a cost-effective application of a PLC system.

1. INTRODUCTION

A variety of applications make use of power line carrier (PLC) as communication channel. These implementations include computer communications, security monitoring and environmental management of buildings. These applications are usually of a relatively small scale, allowing the cost effectiveness of the system to have a lower priority. If one is considering a system that may form the basis of a domestic electricity user's tariffication system, cost effectiveness is of utmost importance.

The main requirement of the proposed system, is that of real time data. The administration board requires a breakdown of a particular user's energy consumption for a given period of time. The length of the measurement period is variable at this stage (due to the fact that a time-of-day tariff is still to be finalised).

Our South African energy supply situation has other peculiarities that need to be considered. The fact that a large percentage of the lower income group currently has no electricity supply is a problem that needs to be addressed. The concept of "prepaid power" is also not a new one. Again a cost-effective solution for the administration of prepaid energy supply is required.

Having set out the requirement, that a time-of-day tariff (for domestic users) is still to be finalised, we realise that extensive research in the area of domestic energy usage is required. The infrastructure for such research includes a data logger to gather the research material.

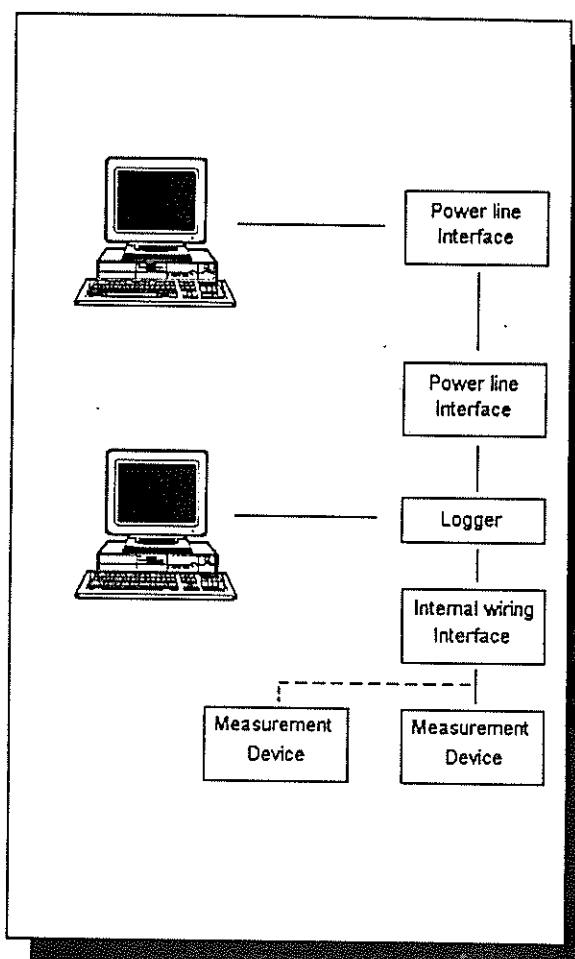


Figure 1 General layout showing the various components

Bearing these facts in mind, a project to fulfil these needs is presented.

II. SYSTEM OUTLINE

The system may be broadly divided into six subsystems (as outlined in figure 1):

Central computer: Located at a mini substation, this computer contacts the data loggers (described hereafter), and downloads a data package. This data (for each user on the particular branch of the distribution network) is then made available to the supply authority via a dial-up modem.

Power line interface: This interface is the communication link between the central computer and the data loggers (located at the distribution point at each user's premises). The energy distribution lines are used as communications medium (ie PLC is employed).

Data loggers: Each user has one of these units connected to the main energy supply lines to his premises. Energy usage data is obtained from energy measurement devices (described hereafter), and stored in onboard non volatile memory. A further communication port allows a local computer to be connected.

Local computer: A local personal computer may be connected to the data logger to facilitate datalogging in the absence of a complete installation (central computer and power line interface not used).

Internal wiring interface: An interface that allows the data loggers to communicate with special energy measurement devices is provided. The internal wiring of the user's premises is used as the communication medium.

Energy measurement devices: Self contained energy measurement devices may be connected to any of the user's electrical equipment. These devices communicate via the internal wiring interface with the data logger. Other than the energy measurement data, a further feature allows the particular piece of equipment connected to the device to be switched on or off under instruction from the data logger (in turn receiving its instruction from either the local computer or the central computer). In the final implementation of this device, only one will be used, and it will measure the total energy consumption of a given user.

III. DETAILED DESCRIPTIONS

Detailed descriptions are only provided for those

devices that were specifically developed for this project - viz.: the power line interface, the data logger, the internal wiring interface and the energy measurement modules. Descriptions of the software required by the personal computers will be presented in the relevant sections.

3.1 The power line interface: Due to the complex and dynamic nature of the energy distribution network impedance, the choice of Pseudonoise (PN) modulation was made. By nature PN (Spread spectrum) signalling provides immunity to narrow-band signal degradation, and allows a high degree of security.

PN signalling spreads narrow-band data signals over a wide band using a pseudorandom code. During reception the same code used during transmission is required to recover the data. By using unique codes, protection from unauthorised eavesdropping, and the facility to multiplex data is provided.

Hardware synchronization problems associated with PN signalling [1] are overcome by using the 50 Hz zero crossings of the power frequency as synchronization pulses. Block diagrams for the transmitter and receiver are presented in figure 2.

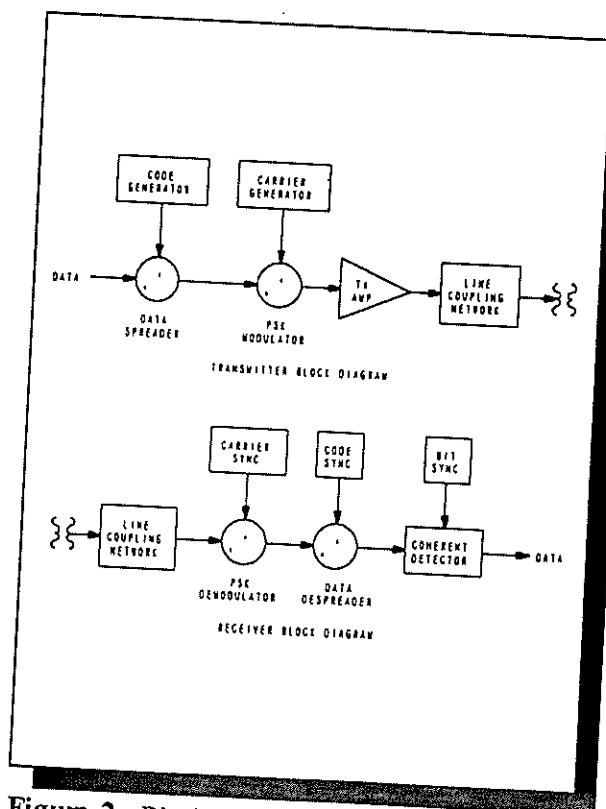


Figure 2 Block Diagrams for the power line interface.

Problems associated with using this modulation technique are solved by the correct choice of PN code sequence frequency [2]. By differentially encoding the data, polarity considerations are obviated (connection to the power lines). The power line interface forms a transparent link between the central computer and the data logger (typically at a rate of 300 bits per second).

3.2 The data logger: The data logger has as communication partners: the central computer, the local computer, and up to sixteen energy measurement devices. This sets the requirement for at least three serial communication ports. The power line interface provides a transparent digital link with the central computer. Each logger has an address code to allow multiple loggers on the same branch of the distribution network. The local computer is connected via an RS232 link. The internal wiring interface provides the link with the energy measurement devices.

The data logger consists of the three serial ports, some non volatile memory (2048 bytes) and a battery backed real time clock.

The data logger operates in one of three possible modes: research mode, time-of-day mode and prepaid mode. Commands are received from one of the two computers, and the corresponding mode is entered.

In research mode, the logger poles the energy measurement devices (up to sixteen of these) on a per minute basis. The energy measurements are stored in a cumulative, or time-of-day fashion (user selected). The controlled equipment may be switched on or off under control of one of the two personal computers. The two personal computers have software that communicates with the logger to download the relevant energy usage information.

In time-of-day mode, the entire energy usage of the user is logged at certain time intervals, and the data is downloaded by the central computer. In this case only one energy measurement device is required.

The prepaid mode makes energy available to the user as long as the energy has been paid for. Once the prepaid amount of energy has been used, the connection to the distribution network is broken. The prepaid amount is dispatched to the logger from the central computer, and in this case, only one energy measurement device is used.

3.3 The internal wiring interface: This interface performs the communication between the logger and the energy measurement devices. Cost effectiveness was the main emphasis in the design of this interface (each user may require a number of these devices).

The lower range, and lower bit tempo dictated the use of a simpler modulation technique. The final choice was amplitude shift keying (ASK). This technique allows very easy demodulation of the data using an envelope detector.

The internal wiring is used for two types of communication: communication from the logger to the energy measurement device, and communication from the energy measurement device to the logger. Rather than making use of a data coding technique to identify the type of communication, a modulation technique was used. The communication from the data logger to the energy measurement device has a carrier frequency of 300 kHz and the communication from the energy measurement units to the data logger have a carrier frequency of 100 kHz. The energy measurement units ignore the transactions that other energy measurement units may be having with the data logger.

The data logger is connected to a free standing internal wiring interface via a RS232 type serial link. The internal wiring interface is connected to the internal wiring at an electricity outlet. A similar line coupling network as that used in the power line interface is employed.

3.4 The energy measurement devices: The energy measurement devices make use of low cost microcontrollers (Microchip PIC Series). The communication interface is incorporated into the unit.

An energy measurement component manufactured by SAMES is used. This device is coupled in line with the appliance, and the energy usage of the appliance is measured. The SAMES device provides a frequency output proportional to the energy usage being measured. The microcontroller measures the frequency, and calculates the corresponding energy.

Communication is controlled by the logger, and the microcontroller in the energy measurement device facilitates the communication. Control of the appliance energy supply is done by a relay, or solid state switching element. The energy measurement device accumulates the energy usage until the logger downloads the latest reading, after which the current

energy measurement is reset, and restarted.

Where the total energy of the user is being measured, a different method of energy measurement is employed (due to the higher energy).

III. IN USE

Due to time constraints, it was not possible to implement and evaluate all the functions of the system. However, the research system as outlined above is complete, and only requires consolidation into a working system. The features of the data logger in the research mode of operation are complete, but the features of the prepaid system still require development. The communication side of the project has delivered satisfactory results, and will prove its integrity in the field.

IV. CONCLUSION

The project set out to prove that a cost-effective form of domestic energy measurement and control can be realised with the use of a PLC system. The system outlined here, can meet all the requirements originally set out. However, further work is needed to achieve all the features discussed.

Future uses of the system, may allow the replacement of the existing "ripple control" systems (used to control domestic geysers), used by many authorities. It will also reduce the amount of labour required in the "meter reading" process. With the flexibility of the recommended system, authorities may exercise a greater degree of measurement and control of the distribution of energy.

It is our belief that the research done in this project may contribute to the more efficient use of electricity in the years to come.

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EVALUATING THE SYSTEM IMPACT OF DECENTRALISED DOMESTIC ENERGY CONTROLLERS

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ABSTRACT

A multitasking simulation environment for the evaluation of the system impact of decentralised domestic energy controllers is proposed in this paper. The proposed software is employed to generate usage patterns, to simulate controller response to domestic device behaviour, and to calculate resultant electrical usage patterns. Multiple cases are then added together in order to arrive at the system impact. These methods all form components of a process termed Computational Load Research (CLR) which enables load research to be performed by computer.

1 INTRODUCTION

1.1 Background

Demand-side management is the process which an electricity supplier follows in order to obtain desired customer load levels and consumption shapes (load curves). DSM strives to improve the total cost-effectiveness of the electricity supply and its consumption. If future and existing demand profiles can be optimised and their shapes improved, significant cost savings may be brought about [17]. The research work being discussed in this paper falls directly into this category and is aimed at controlling individual loads according to specific demand-side management criteria. One of the major components of performing demand side-management, is the gathering, manipulating, simulating, generating and interpreting of load data. This process, which has become highly automated and makes almost exclusive use of modern computer technology has been termed "Computational Load Research" [12].

1.2 The Importance of the Domestic Sector

The domestic sector has the largest number of energy users in the Republic of South Africa. Although their combined electrical energy component is estimated at 14 %, the domestic load is the largest contributor to the peak of the national load [17]. High morning and evening peak loads, caused by the domestic sector, are of great concern to electricity suppliers and therefore represents a very important field of research [3].

1.3 Decentralised Controllers

Decentralised controllers offer the end-user the advantage of having control customised to his specific requirements. It is proposed that they also promote additional cost and energy savings and offer additional comfort enhancing features. It has also become evident from research that a significant modification of the system load profile results from the use of intelligent decentralised control schemes. In order to design, evaluate, modify and enhance these decentralised control structures, a specific set of software tools are required [4,5,6]. A hardware exercise would prove impossible, given the extents of the options to be tested and thus CLR is employed.

2 IMPLEMENTATION

Software modules performing the different tasks were all developed according to an object oriented approach. Through the software mechanisms of inheritance and encapsulation, complex coding structures become significantly simplified and greatly encourage a standard library of usage pattern generators, physical device models and controller structures [16]. Using object oriented programming to realise CLR, has lead to greater productivity and reliability. This is so because previously tested and completed code could now be incorporated quickly and easily into new simulations [8]. OOP is an approach to software engineering that has therefore greatly influenced the work being described in this paper.

In order to start the simulation process, independent device usage patterns are generated by probability density functions which all have practically measured deviate distributions [9,10]. The specific controller algorithm under investigation, is then fed with these software generated usage patterns and further responds to physical device models, in order to generate the required electrical consumption patterns [11]. External pricing signals as well as environmental and customer social pattern data can also be included as inputs to the simulation process, and the combined software generates one electrical consumption pattern per house. The electrical usage

pattern of N of these occurrences (e.g. 50 houses with pool-pump timers) are subsequently added together in order to produce the electrical totalized electrical consumption pattern. All the simulations are performed using hard-number-crunching algorithms, running on a 50-MHz-486 and therefore avoid statistical averaging assumptions in order to arrive at results.

3 FUZZY LOGIC CONTROL

The development of fuzzy logic controllers allows complex control requirements to be implemented in simple, easily maintained, and inexpensive hardware. Fuzzy logic is a methodology of easily representing analog processes on a digital computer. These processes are concerned with continuous phenomena that are not easily broken down into discrete segments, and fuzzy excels when the concepts involved are difficult to model along mathematical or rulebased lines [2].

A good example here might be the hot water heater controller for a typical home. The control rules could include the water consumption used or expected, the current tank temperature, and the current price of electricity. Input variables then have ranges of states in such a control system, e.g. temperature could be "very hot", "hot", or "slightly hot". The ideas of what is "hot" and what is "very hot", have different interpretations in different situations, and such conditions combined with a highly non-linear physical device such as a geyser, lends itself perfectly to fuzzy control structures.

The fuzzy controllers being used in the author's work employ fuzzy associative memory matrixes which are set-up and implemented in software [13]. Any specific FAM matrix, together with its associated membership functions for the input and output variables, and their scaling factors, are then down-loaded into the actual controller configurations. A typical practical fuzzy controller result, for performing domestic maximum demand control, is demonstrated in the results section.

4 NEURAL NETWORK CONTROL

Neural networks form a significant portion of the CLR being performed on domestic decentralised control systems. There are several reasons for this but probably most importantly is the fact that these systems are of themselves adaptive structures. These controllers can learn from presented data and can therefore learn from experience. Previously unknown solutions may be inferred, avoiding the condition that exact knowledge about the system under investigation

be known beforehand (ideal for learning customer usage patterns or predicting load curve behaviour) [15].

Neural networks can generalise, which means they exhibit fault tolerance and can process imperfect or incomplete data. This is ideal for "noisy" real world control implementations. Thirdly, neural controllers are highly non-linear and can capture very complex interactions among the input variables of a system. The relations between for example price, comfort, energy savings, and preventative control is an example of such a complex interaction and can be learnt by a typical network [1,14].

5 ELECTRONIC CONTROL

Rapid advances in compact digital electronics have enabled software controlled "smart controllers" to become technologically and commercially viable [18,19,20,21]. The major features of these systems has been their ability to perform time differentiated control (i.e. type of electronic timers), they have communications channels by which external devices can be switched on or off or signals can be relayed to them and they can implement software based control algorithms. These systems however lack the "intelligent" nature of the systems proposed in this paper, but have already found their way into the commercial market. These systems should however also be evaluated both from the consumer's point of view (e.g. a cold shower, energy savings), as well as from the system side, where their combined electricity consumption profiles should also be superimposed on the existing load profiles to determine their group response.

6 PRACTICAL RESULTS

6.1 Verifying the Technique

In order to validate the proposed technique and to test the numerical accuracy of the software being employed, the measured geyser consumption patterns of 7 arbitrary homes situated in Hatfield, was compared to a result generated by simulation. In order to achieve these results, statistical methods were employed to generate estimated water consumption patterns and these were subsequently fed into physically based load models for geysers (solutions of sets of differential equations) and these simulated geysers were then controlled by software implemented thermostats.

The results of simulated versus practical results showed a very good correlation for as few as 7

consumers and these results are displayed in Fig. 1 and Fig. 2. respectively.

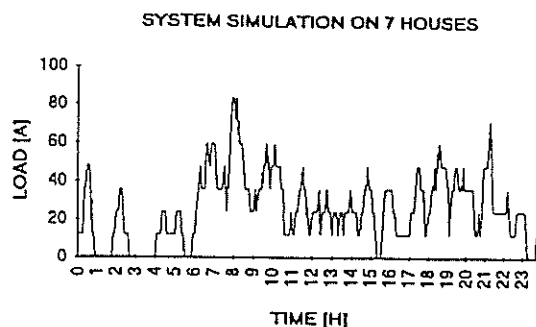


Fig. 1. Simulated geyser consumption for 7 houses.

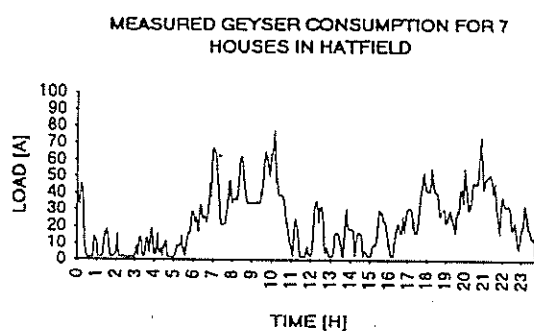


Fig. 2. Practically measured geyser consumption for 7 houses

6.2 Centralised Control Simulation

The developed software also allows centralised control options to be investigated and a geyser consumption pattern for 50 houses, with and without a centralised control signal is shown here. The block being simulated was switched off between 09h00 and 10h30 and the results are shown in Fig. 3 and Fig. 4. respectively. The "cold-load pickup effect", which can nullify the original peak-cutting intention, is clearly visible in this result. This type of simulation also allows the effects of element sizes, tank sizes, inlet temperatures and thermostat settings to be evaluated.

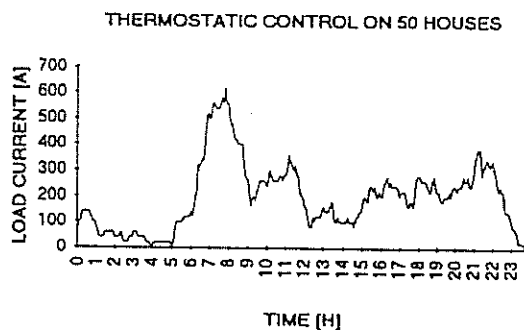


Fig. 3. System response of 50 houses.

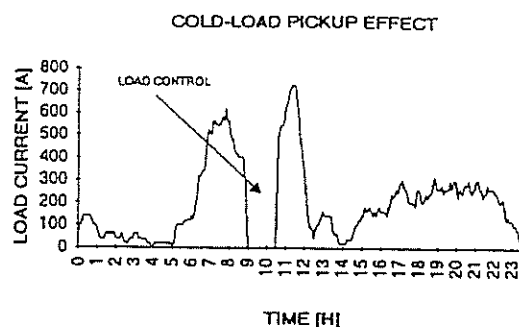


Fig. 4. System response of 50 geysers with an interrupted-supply control signal.

6.2 Fuzzy Logic Controller Implementation

In order to illustrate the possible effect of even very elementary fuzzy logic control systems, a two input variable, fuzzy logic controller was installed in each of 20 geyser simulations on 20 houses. These fuzzy logic controllers measured temperature, as well as the total block load current (transmitted from a common feed point) and the rules were constructed so as to limit total load current to the vicinity of 100 Amps. The limiting function however was "soft", as it also depends on the current tank temperature via a fuzzy relation. The results shown in Fig. 5. and Fig. 6 show that the fuzzy controllers avoid the high system peak and rather spread out the required energy into those areas where group consumption was lower.

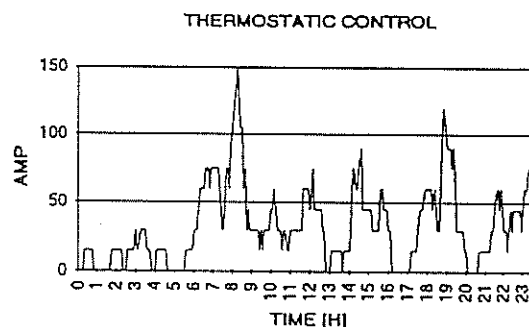


Fig. 5. Electrical consumption for 20 houses.

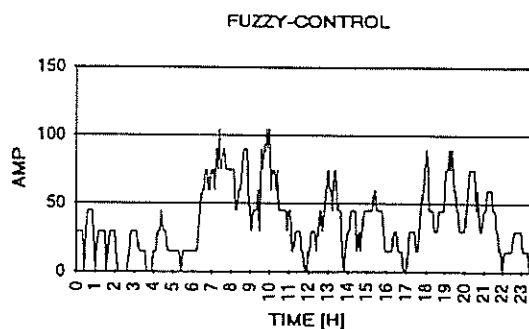


Fig. 6. Electrical consumption for the same 20 houses using fuzzy logic controllers.

7 CONCLUSION

This paper described the Computational Load Research which is being performed by the authors in the field of decentralised domestic energy controllers. A technique for evaluating system impact results was addressed specifically. The various implementations of the intelligent devices being developed are proving that the following options are practically realisable :

- i) performing energy cost savings on time-differentiated tariffs,
- ii) preventing unwanted group response behaviour; provided communication channels are available (e.g. cold-load pickup effect),
- iii) acting as a domestic "maximum demand controller",
- iv) performing preventative control options (e.g. preheating water to a higher than normal temperature to make provision for expected excessive hot-water consumption),
- v) performing energy savings by minimising losses and eliminating energy wastage,
- vi) performing additional utility tasks such as pool-pump control, washing-machine cycling, security light control and domestic metering.

Future work will entail formulating the optimum combination of the structures discussed in this paper, and evaluating and enhancing its responses to various sets of customised requirements.

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THE ROLE OF DOMESTIC ACTIVITY AND COMMODITY CONSUMPTION AUDITS IN DEMAND RESPONSE MODELLING

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ABSTRACT

Customer behaviour and preferences probably have the largest influence on the aggregated load profile of the residential sector. Physical attributes of domestic appliances and the way we construct our home also has a significant impact. If we wish to improve energy efficiency and reduce electricity costs to the home, it is necessary to gain a better understanding of the potential domestic demand response to cost reflective (dynamic) electricity tariffs.

A procedure for collecting the necessary information via telephone audits and cost effective measurements is proposed. The analysis of this data and the estimation of parameters for load models is described. Finally, it is shown how domestic activity and commodity consumption audits can be used to estimate the potential demand response to time-based tariff structures.

1 INTRODUCTION

In order to successfully develop accurate load models to facilitate demand side management of household loads, accurate information of typical usage patterns is required. The studies presented here aim to help establish the required data. One method concentrates on measuring the hot water demand of the household in the most cost effective way. The second approach aims to establish typical load profiles of the remaining major energy consumers in the home.

2 TELEPHONE AUDIT

2.1 Indirect energy Audits

The audit strategy proposed here involves the asking of certain questions in order to obtain the necessary data. This type of audit has some inherent problems. These problems originate because people have different perceptions with

regards to the same situation. It is up to the designer of the questionnaire to devise questions which minimize the variance in outcome. A further problem is the accuracy of time of use data. Residents generally know the time they make use of a certain appliance to within a tolerance of half an hour. This determines the level of accuracy of the questions that are posed.

2.2 Obtainability of data

All the uncertainty implicit in the use of this type of audit raises the question of which type of data can be successfully obtained.

a)Appliance penetration. This is perhaps the most easily obtainable data through the use of questionnaires. Residents simply state the number(s) and type(s) of appliance(s) that they own.

b)Power classification.

It is almost as easy to obtain the power rating of the different appliances (usually information appears on the back of the appliance). Items such as stoves and ovens pose a bit of a challenge, but manufacturers data sheets help to establish a workable average rating.

c)Classification data.

Through the use of selected (carefully chosen) questions classification data such as type of house, type of household, approximate income, etc. may be obtained.

d)Time-of-use information. This information is essential in order to apply the model. However, care should be taken not to overestimate the accuracy of information obtained in this way (plus or minus half an hour tolerance has been found to be realistic).

e)Consumer attitude.

This is another category of data which can be successfully collected through the use of questionnaires. It is this aspect in particular which enables us to estimate the price elasticity of demand.

2.3 Design of questionnaire

When drawing up the audit form careful attention must be given to the selection of questions, as well as to formulation of these questions. It is important to continuously keep the respondent in mind. Minimum technical detail should be included. Notwithstanding this, it should be noted that the outcome of the study will be greatly influenced by the quality of the data obtained. Questions should be formulated to extract the maximum data from the respondent, in the shortest possible time.

2.3.1 Categories

To compose and draw up the required questionnaire, seven categories of energy consumers in the home have been identified: [1]

- a) Water heating (Geysers)
- b) Lights
- c) Food preparation
- d) Space Heating
- e) Pumps
- f) Washing (clothes, dishes)
- g) Cold storage

The emphasis of the telephone audit was placed on categories b through e ie. Lights, Food preparation, Space Heating and Pumps.

3 DATA PROCESSING

3.1 Use of computer

In order to optimise the auditing process, the use of a computer was a natural choice. A data managing program was written in Visual Basic. Visual Basic is a Windows-based programming tool. By using Visual Basic, the user interface is the familiar Windows graphical interface. The program asks a series of questions, which the operator repeats to the respondent. Data is stored in such a way that it may be retrieved and altered at a later stage. Data processing is also performed by this program.

3.2 Model

The mathematical model being used to process the data is very elementary. This is desirable however, since the model is not expected to modify or manipulate the results of the audit in any way. The

model should simply give an undistorted reflection of the audit. A certain amount of "smoothing" of data is built in, due to the way questions are formulated. The model used is that of equation 1.

$$P(t) = \sum_j^m P_j * N_j * f(t) \quad \dots \quad (1)$$

Where:

- $P(t)$ = total power as function of time
- P_j = power consumption of appliance j
- N_j = number of appliance j
- m = different types of appliance j
- $f(t)$ = usage function

The usage function $f(t)$ is generated by using data supplied by the audit and the auditing program. The usage function of each appliance is true when the appliance is switched on (value = 1), and false when the appliance is switched off (value = 0).

4 RESULTS

4.1 Generating graphs

The output of the audit program is in a suitable format to allow analysis by Lodan. Lodan is a load analysis program developed by Cnes (The Centre for New Electricity Studies in the Department of Electrical and Electronic Engineering at the University of Pretoria). The drawing of different load curves (i.e. load profile, stacked load profile, percentage load plot etc.) is made simple by using Lodan.

4.2 Interpretation of graphs

When looking at the graphs obtained by the use of this method, it is important to realize the restrictions imposed by the system. The graphs do not reflect an absolute truth, as the data is viewed for an "average weekday", as interpreted by the respondents. Furthermore it should be realised that not all consumers of electricity within a specified category are portrayed, only the main consumers are represented. Provided enough thought has been given to the planning and design of the questionnaire, results should prove to be satisfactory. A number of usable deductions can be made, these include:

- a) General form of load profile
- b) Load profile of individual categories.
- c) Dissaggregated load profile
- d) Identification of problem categories.
- e) Possibilities for demand control and load rescheduling.
- f) Probable response to variation in energy tariffs.

g) Price elasticity of demand.

All graphs included in this paper apply to an audit performed on the suburb of Lynnwood in Pretoria during May 1993. The data was obtained from twenty houses.

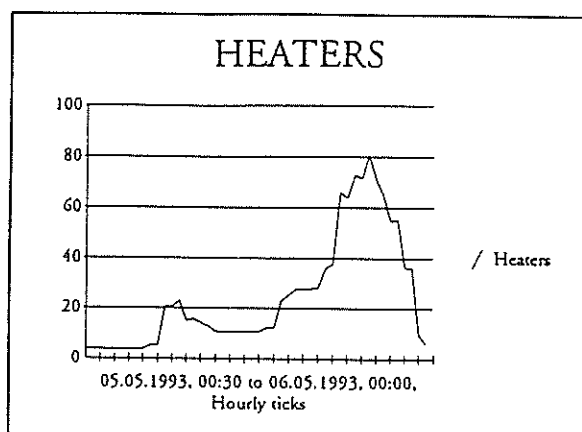


Figure 1 (KW vs Time)

Figure 1 represents the load profile of space heating appliances. The well defined peak in the evening and the much smaller one in the morning, give a clear representation of the way these appliances are utilized. In order to see the relative contributions of four of the largest energy consumers (hot water load excluded), a stacked load profile is shown in figure 2.

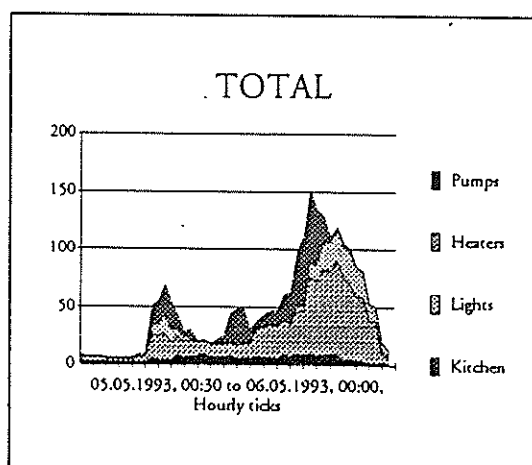


Figure 2 (KW vs Time)

From figure 2 the effects of the different categories can easily be seen. By analysing this data, it was found that the load profile of pumps generally matched the total load curve. Immediate identification of demand site control opportunity is

possible. By switching off pumps during peak hours the total demand may be reduced by as much as 10 KW (20 houses). This translates to a reduction of 14% in the morning and nearly 7% in the evening. Residents were asked if they would be willing to change the times pumps operate, if this would bring about a smaller electricity bill. The response was generally positive, nobody was interested in changing the pattern without the motivation of savings however. We believe that further studies of this nature would give a good indication of the potential price elasticity of demand.

5 COST EFFECTIVE ENERGY AUDITS ON ELECTRICAL WATER STORAGE HEATERS

A portable system for measuring hot water consumption patterns in households by means of temperature- and input power measurements is being developed. To be cost-effective it does not use any intrusive flow meters and must be conveniently installed in any household to measure the water consumption patterns and the energy efficiency.

The method proposed, makes use of only two temperature sensors. The two sensors are fastened on the outlet pipe of the geyser with a distance of about one to two meters between them. They should be as near to the geyser as possible without being affected by the losses from the geyser through the first couple of centimetres of the pipe that is always hot. The reason why two sensors are used, is to find a way to determine the rate of water flow through the pipe. If the length (in m) between the sensors (L), the inside radius of the pipe (r) and the difference in time (δt) at which the temperature starts to rise at the two sensors are known, the rate of water flow inside the pipe can be calculated with the following equation:

$$\text{Flow rate} = 1000 \cdot \pi \cdot r^2 \cdot (L / \delta t)$$

The data-logger being used, is a XT computer with a PC30 card on board.

The reason why a computer is used is simply because there is so much data to be stored on the hard drive and a data-logger that has enough storage space couldn't be found, or is too expensive. The sensors are coupled through the right connections with the computer (depending on the type of sensor). A connection is made from the element of the geyser to one of the ports of the PC30 card. This is necessary in order to measure the power that is being put into the system.

The data logging is done through a software program written in Turbo Pascal. The program records the temperature of both sensors and the voltage over the element every 100ms. The reason for this relative high sampling rate is to apply the equation for flow rate effectively and to get useful results from the data. The program measures for three minutes and if something has happened, it will store the time at the beginning of the measuring interval at the top of the file followed by all the values that have been measured. The data can then be analyzed.

It is possible to calculate when water was retrieved from the geyser because of the temperature rise on the outlet pipe. The exact amount of power that was put into the geyser can also be seen because the program stores the data if the element was on in a measuring interval. The rate of water flow through the pipe when water has flowed can be calculated by using the equation for flow rate.

There are, however, a few problems with this method.

The measuring system was installed inside a home for experimental purposes. The main problem encountered was that the pipes in the home were thick galvanized pipes and not copper pipes and the heat transfer to the temperature sensors was not satisfactory.

However, because the input energy to the system, the relative length of time that the tap was open, the time when water was used and the relative flow rate is known, careful analysis of the data can be done and by comparing the input energy to the output energy, valuable, usable information can be obtained by this method.

6 CONCLUSION

The methods developed will improve our ability to produce valuable information to help gain a better understanding of the load profile and usage patterns of the domestic sector. By applying these techniques correctly it may be possible to investigate aspects such as price elasticity of demand and consumer preferences. These in turn can help the development of successful time-of-use tariffs.

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The use of a voltage window to manage noise in partial discharge measurements

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1 Abstract

This paper proposes that the injection of an impulse, at a suitable time on a power frequency voltage wave, can trigger the inception of partial discharges. In this way a deterministic component is introduced into the observation of partial discharge activity and the ability to minimise the influence of noise in a measurement is considerably enhanced.

Two simple experiments are described which demonstrate the triggered inception of partial discharges and the strong "focusing" effect of the triggering impulse, whereby the discharge activity is concentrated immediately after the injection of the impulse and a significant "quiet" period follows in which minimal discharge activity occurs.

2 Introduction

Partial discharges occur when the electric field in a gas filled cavity in solid or liquid insulation system reaches the breakdown value for the gas. This discharge causes the voltage across the cavity to collapse and the voltage across the remainder of the insulation to rise by an equal amount. The redistribution of charge which results gives rise to a very short current pulse in the supply leads. The typical duration of this pulse is 5-20ns [1] and the charge is small, being typically in the 10-1000pC range. With the low detection impedances necessary to minimise noise, the observation of partial discharges requires the extraction of very small voltages from many tens of kV. Mains borne and radiated noise can readily swamp partial discharge signals and mains filters and fully screened test areas are required for sensitive measurements.

With the increasing demand to do partial discharge measurements on site, where screening is not feasible, elaborate noise control strategies have been explored. Discrimination techniques [2], whereby observations are excluded if they coincide with noise, suffer with the problem that if the partial discharges occur at the same time as noise they are excluded from the observations. Post-observation signal processing relies on the ability to characterise the noise at reduced voltage and then "subtract the noise" at the test voltage [3]. For on-line measurements the computing overheads for such processing are very significant and benefit

would be gained if some means could be used to improve the signal-to-noise ratio of the input signal.

This paper investigates the use in an impulse, synchronously superimposed on the ac test waveform, to initiate partial discharge activity at selected positions in the cycle. If the noise has deterministic aspects, the partial discharge activity can be initiated in "quiet" zones. If the noise is stochastic in nature, the knowledge that the discharge occurs in a limited time window enhances the total signal-to-noise ratio. Further, the narrower the window in which the partial discharge is known to occur, the greater the signal-to-noise ratio and the more meaningful the measurement.

The concept of initiating a partial discharge by means of an externally impressed lightning impulse, superimposed on the power frequency waveform, has been used for the analysis of PD inception levels in oil-insulated equipment [4]. The superposition of switching surges on the power frequency waveform in laboratory tests with oil-insulated equipment has been used to aid the characterisation of PD activity [5]. However, in neither case has either synchronicity with the power frequency waveform or repetitiveness in successive cycles been investigated. Both of these factors are essential features in noise control strategies explored in this paper.

3 The injected impulse

The very short duration of the partial discharge current indicates that the breakdown and conduction processes in a gas filled cavity are very rapid. Some uncertainty exists regarding the statistical time lag [6], which is related to the combination of the properties of the insulating material and the residual space charge remaining from discharge activity in previous cycles. The rise time of the injected impulse has to be significantly slower than that of the expected partial discharge so that it can be readily distinguished from the partial discharges. It also has to have a duration which is sufficiently long to ensure that there is adequate time for all the breakdown and conduction mechanisms to take place, whilst being as short as possible for effective noise control. Bearing these constraints in mind, a somewhat arbitrary choice of an impulse was made. The rise time was 400ns and

the time to decay to zero was $85\mu\text{s}$. Some overshoot followed. The waveform is shown Figure 1.

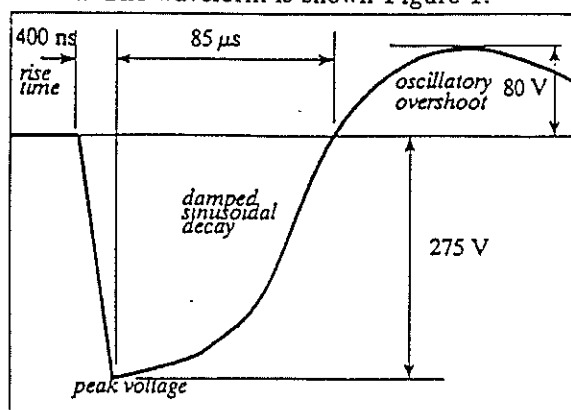


Figure 1. Injected impulse

4 Experimental work

4.1 Apparatus used

The equipment was arranged as shown in Figure 2 and the details of the test sample and electrodes are shown in Figure 3. The sample consists of two layers of polyethylene as a tightly compressed sandwich, of which one layer had a 4mm diameter hole punched in it. The hole was centred over the detection electrode whilst the guard electrode was grounded.

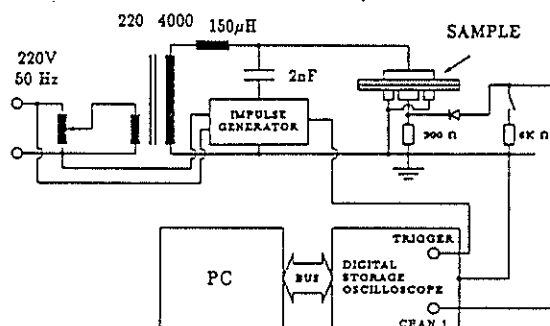


Figure 2. Major components in the experimental setup.

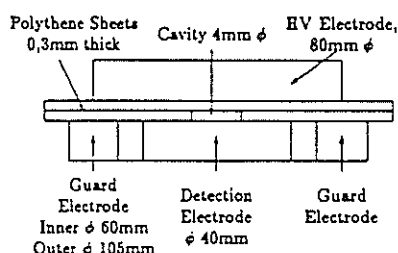


Figure 3. Electrode and sample configuration

For the dimensions used, partial discharge inception and extinction was about 1 150V and 900V rms

respectively. After taking control results for comparison purposes, the impulse generator was introduced into the ground end of the parallel capacitor. This capacitor performed the dual functions of high frequency bypass capacitor for partial discharges and coupling capacitor for injection into the high voltage end of the sample.

The output of the detection electrode was grounded by a 200Ω resistor. Across this resistor was a crude negative-polarity peak-reading pulse stretcher circuit with two values of decay time constant available (viz. approx $2\mu\text{s}$ with $6\text{k}\Omega$ in circuit and $80\mu\text{s}$ with only the $1\text{M}\Omega$ and the capacitive input impedance of the digital storage oscilloscope). Due to the fact that the most significant discharges produced some volts across the grounding resistor, the error in the generation of effective peak voltage across the resistor, introduced by the diode, was not considered serious for the primary objectives of the experiment.

Data capture of the output was achieved using a digital storage oscilloscope with sampling rates ranging between 500kSa/s and 5MSa/s . The record length was 16kbytes

4.2 Experiment 1

The power frequency excitation waveform was adjusted to slightly below inception voltage, and the injected impulse was applied with reference to zero-crossing into the negative half-cycle. Beginning at zero delay from the zero-crossing reference point, the delay of impulse application was incremented in 1ms steps. No discharge activity was recorded for 0, 1, 2 or 3 ms whilst activity was observed for delays of 4 and 5ms. On regressing in 1ms steps towards zero delay, hysteresis was noticed as activity continued at 3ms. Figure 4 illustrates a typical observation.

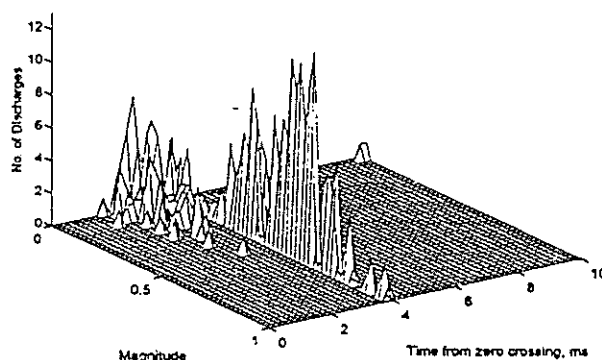


Figure 4. Partial discharge activity with the injected impulse delay advanced to 4ms, 900V rms power frequency, 200 measurement cycles. The partial discharge magnitude has been normalised to approximately 100pC

4.3 Experiment 2

The power frequency excitation voltage was increased to above inception to guarantee discharge activity.

Upon inspection of the discharge distribution without the injected impulse, a delay of 2ms was arbitrarily selected to offer a narrow window, 1,6ms wide, enclosing substantial normal activity. Figures 5 and 6 show the activity distributions without and with the presence of the injected impulse. Figures 7 and 8 are expanded views of the distributions for 50 μ s after the injection of the impulse. Figure 8 is a rotated version of Figure 7 to illustrate the very rapid decay of activity after the injection of the impulse. These distributions underscore very strongly the concentrating effect of the injected impulse.

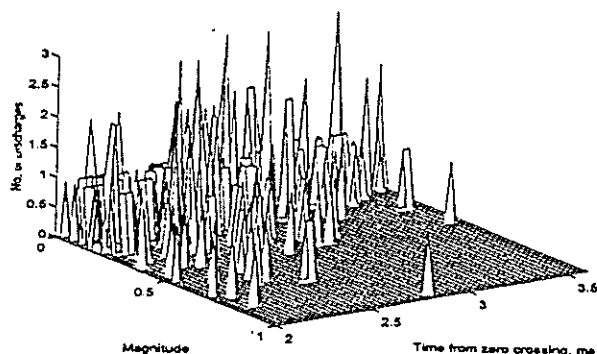


Figure 5. Partial discharge activity, in a 1,6ms window, 2ms after zero crossing without an injected impulse, 1050V rms power frequency, 200 measurement cycles. The partial discharge magnitude has been normalised to approximately 100pC

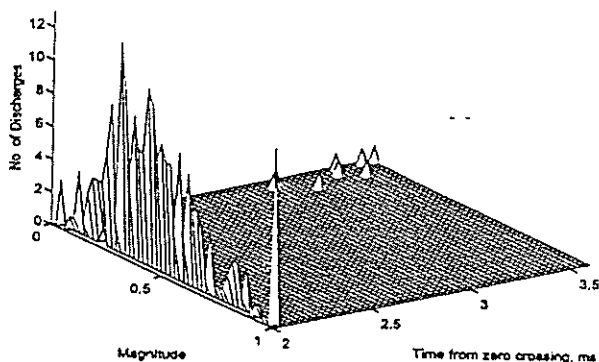


Figure 6. Partial discharge activity, in a 1,6ms window, 2,0ms after zero crossing with an injected impulse having a 2,0ms delay, 1050V rms power frequency, 200 measurement cycles. The partial discharge magnitude has been normalised to approximately 100pC

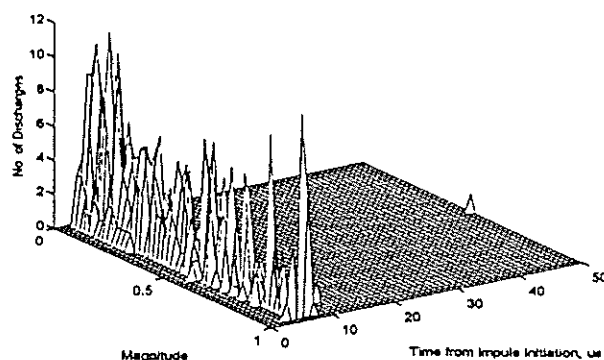


Figure 7. Partial discharge activity, in a 50 μ s window, 2,000ms after zero crossing with an injected impulse having a 2,000ms delay, 1 050V rms, 500 measurement cycles. The partial discharge magnitude is normalised to approximately 100pC

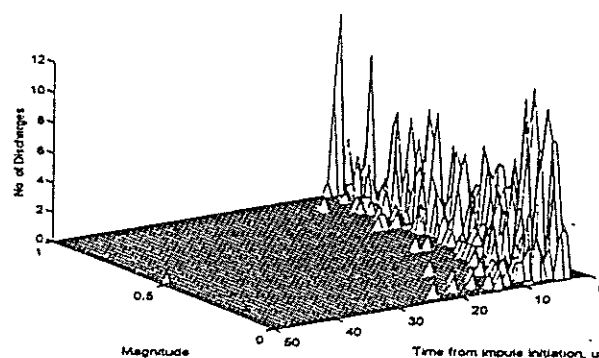


Figure 8. Rotated view of partial discharge activity, in a 50 μ s window, 2,00ms after zero crossing with an injected impulse having a 2,00ms delay, 1 050V rms, 500 measurement cycles. The partial discharge magnitude is normalised to approximately 100pC.

5 Comment on experimental work

From the experimental work, the following observations can be made:

5.1 Experiment 1

1. Injection of the impulse on the power frequency wave at an appropriate time (approximated by the time at which the sum of the peak pulse voltage and power frequency voltage reach inception levels) initiated discharge activity at levels of the power frequency wave which were normally below discharge inception.
2. Discharge activity was tightly concentrated following the injection of the impulse.
3. A "quiet" period followed the concentration of the discharge activity.

4. The injection of the impulse on only one half-cycle caused asymmetrical cavity behaviour, resulting in the loosely grouped cluster of discharges prior to impulse injection.
5. Isolated low-level discharges after the "quiet" zone could be erroneous measurements derived from rectification of the negative oscillatory overshoots of large positive discharges in the opposite half-cycle. (It was noted that the rise time of discharge signals generated damped resonance in the receiver circuitry.)
6. Hysteresis was observed in discharge activity as the impulse was swept from zero-crossing to the crest of the waveform and back again.

5.2 Experiment 2

1. The injected impulse focused discharge activity very strongly into an extremely narrow window just following it.
2. A substantial "quiet" zone was created behind the impulse.
3. Analysis of the microscopic distribution of discharge activity is related to the sum of time lags for discharge generation. This distribution may provide very provide a useful resource for the identification of the various defects which cause partial discharge activity.

6 Discussion and Conclusions

The fundamental objective of the experiment, namely to attempt to dramatically reduce the range in time over which a particular cavity generates discharge activity under power-frequency excitation, has undoubtedly been achieved. Whereas normally excited discharges were seen to occur over a 2ms range, superimposition of the trigger pulse caused "focusing" of the same activity into a 20 μ s region trailing the application of the trigger pulse. The ability to initiate discharge activity when the exciting waveform is held just below usual inception levels has also been shown.

The impressive concentration of the originally scattered discharge activity, on a macroscopic scale, yields completely the expected result. The distributions observed contain a lot of information which could aid in diagnosis of the discharge source. The physical processes that are operating within the cavity may also be investigated by consideration of the distributions. Theoretically the discharges should be clustered immediately behind the trigger pulse (within less than the rise time) if it were not for physical processes which dictate otherwise.

Although the work carried out is of a somewhat primitive nature, it can be readily seen that the concept of a "voltage window" imposed upon a regular 50Hz excitation waveform may dramatically en-

hance post-event signal processing for the suppression of noise. At an elementary level, the window within which a discharge can be observed under power-frequency excitation has been substantially reduced, thereby reducing the probability of a noise event masking the signal. Further, the window may, with certain restrictions, be moved to an area of the exciting waveform with low noise event count.

Whilst the primary objectives of the experiments were very satisfactorily met, great scope has been opened up for investigation of the mechanisms associated with partial discharge behaviour. The use of the proposed technique as a tool for noise reduction was the original motivation for the work, but it is clear that its application has wide implications for developing a greater understanding of the nature of partial discharge activity and for the development of diagnostic tools.

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"CAPTAP" - A RURAL ELECTRICITY SUPPLY FROM EHV TRANSMISSION LINES

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Eskom

ABSTRACT

This paper describes a novel technique, called CAPTAP, for supplying electricity to communities adjacent to EHV transmission lines. The technique uses the overhead shieldwire on a transmission line to tap small amounts of electricity from the line. In this way Extra High Voltages (EHV) like 400 kV and 765 kV can be stepped down to distribution voltages like 22 kV and 11 kV and used to supply small loads, up to 100 kW. The problems associated with this technique are described and their solutions given. The benefits and advantages of this system are also mentioned.

1. INTRODUCTION

The frustration experienced by rural people living close to EHV power lines and who do not have the benefits of electricity, is a well known problem. A technique for harnessing the power induced in the overhead ground wires has been developed to produce an alternative rural electricity supply. The technique has been given the name CAPTAP, a power TAP using capacitive coupling.

2. PROBLEM DESCRIPTION

It has been known for many years that an electric charge is induced on isolated conductors running parallel to transmission lines. If a continuous section of one of the overhead ground wires, of a transmission line, is isolated (ie. unearthed) from the towers for a certain length, say 10 km, a voltage source can be derived (see Figure 1). This source, however, has a high series impedance which is almost purely capacitive. If the source is loaded with a resistive load the output voltage drops dramatically with increasing load current.

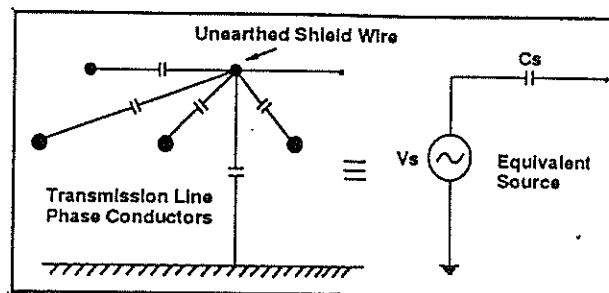


Figure 1 : Equivalent source

3. SOLUTION

There are several techniques for overcoming the voltage regulation problem. A system which is commercially available from overseas, is one which uses power electronic control of a shunt reactor to obtain voltage regulation. On investigation, this system proved very costly and an alternative was investigated.

After numerous Electromagnetic Transients Program (EMTP) computer simulations, ESKOM decided to test the method of series resonant compensation. A prototype was erected on a 400 kV line near Apollo Substation in 1992 to evaluate the method. This involves connecting a fixed inductor in series with the voltage source (see Figure 2(a)). The inductor impedance is selected equal in magnitude to the source impedance at 50 Hz so that the source becomes an "ideal voltage source" at 50 Hz. This gives good voltage regulation under varying load conditions. The phasor diagram in Figure 2(b) shows clearly that the output voltage is independent of the load current.

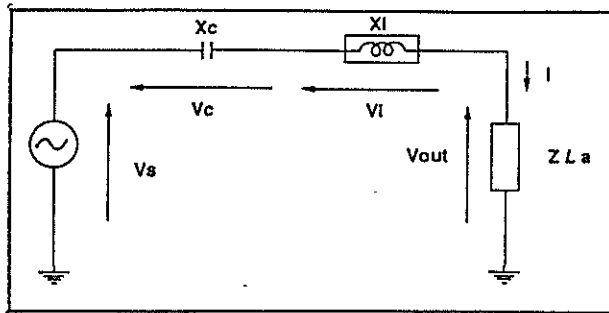


Figure 2(a) : Simplified circuit diagram

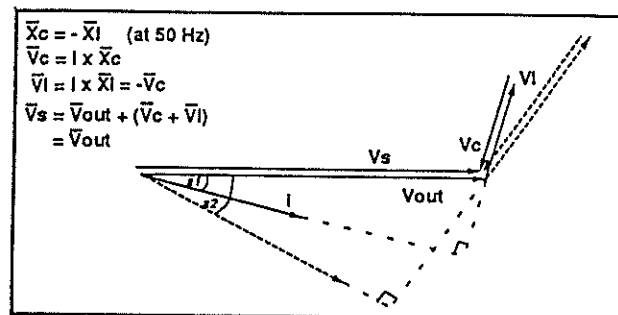


Figure 2(b) : Phasor diagram

4. ADVANTAGES

This technique has the following advantages :

- No electronics needed for voltage regulation
- Simple, robust system
- Low cost
- Reliability due to few components

5. LIGHTNING PERFORMANCE

A question often asked is : "How does insulating the lightning shieldwire affect the lightning performance of the Transmission line"? When the ground wire (lightning shield wire) is isolated from the towers, insulator discs with spark gaps are used on all strain towers. If lightning strikes the shieldwire directly, the spark gaps flash over (at ± 100 kV peak) thus conducting the lightning current through the tower to ground, as it normally does. The unearthed lightning shieldwire maintains its shielding function in this way. A lightning surge arrester is used to protect the CAPTAP system during such occasions.

6. ASSOCIATED PROBLEMS

Inherent in this technique are a number of problems which appear in the practical situation.

6.1. SYSTEM ENERGIZATION

Figure 3 shows the setup of the system when it is to be energized. When the system is energized without a load connected to the distribution transformers secondary side, transient oscillation is set up which causes up to 2 p.u. overvoltages. This is due to the series combination of the source capacitance, the system inductance and the distribution transformer magnetization inductance. The resonant frequency of this combination is excited by the switching "ON" transient.

The damping in this circuit is very small due to large inductance values (45 H) for prototype and low resistance values.

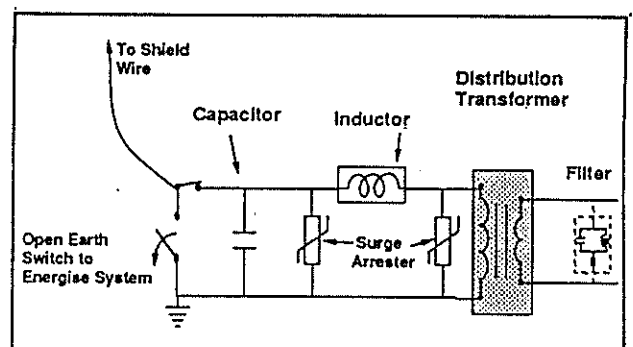


Figure 3 : Schematic diagram of "CAPTAP" system

One solution is to connect a permanent dummy load resistor to damp out this transient energy. This, however, wastes energy during normal "ON" conditions and is not suitable.

A better solution is the application of a parallel resonant filter in series with a resistor on the L.V. side of a distribution transformer. This has the effect of damping transient energies of frequencies other than 50 Hz while not dissipating 50 Hz energy during steady state "ON" conditions.

6.2. LOAD REJECTION

When a large load is switched off, the energy stored in the reactive elements of the circuit oscillates and causes overvoltages in the circuit. This problem is particularly serious when there is almost no load left on the system after load rejection.

The filter used to solve the energisation problem also solves this problem. Figure 4(a) and 4(b) shows the output voltage of the system without and with the filter respectively, for the following events :

$t = 10$ ms Energise system (no load connected)

$t = 200$ ms Switch ON 10 kW load

$t = 250$ ms Switch OFF 10 kW load

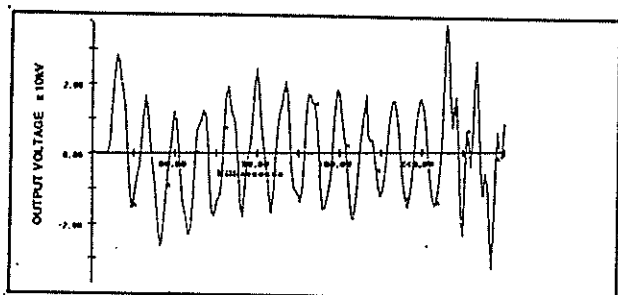


Figure 4(a) : Output voltage without filter

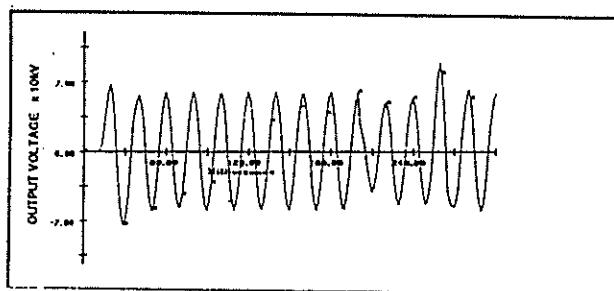


Figure 4 (b) : Output voltage with filter

6.3. HARMONIC GENERATING LOADS

The impedance of the supply is low at 50 Hz, but high at other frequencies. In other words, 50 Hz current can easily be drawn from the source, but harmonic currents do not flow easily from the source.

Loads which generate harmonic currents, such as AC-DC converters, have the effect of severely distorting the output voltage waveform. This is due to the high source impedance at higher frequencies.

On the prototype, a single phase to 3-phase converter was used to drive a 3 phase motor coupled to a 6 kW load. The harmonic currents generated distorted the output voltage virtually into a square wave.

This problem can partly be alleviated by using the filter described earlier. However, this supply is more suited to domestic loads which are mostly resistive, and cannot be expected to handle heavy thyristor controlled loads without voltage distortion.

7. SYSTEM PROTECTION

The nature of this system prescribes that an earthing switch on the shieldwire should be used to switch the system off in case of fault. This is to prevent large overvoltages caused by series switching of inductive currents.

A 12 V based, low cost protection panel is used to protect the system against overcurrents, earth faults, undervoltages and protection component failures. The protection was designed to be a low cost, robust and failsafe system which requires little maintenance, but is also effective in protecting the primary plant in cases of faults.

8. PROTOTYPE PERFORMANCE

A prototype was built near Eskom's Apollo Substation to test the concept of series resonant compensation. Although certain problems were encountered as discussed previously, the steady state performance of the system is promising.

Output Voltage	11 kV.
Max. Power Output	30 kW
Shieldwire Length	10 km
Voltage Regulation (at 11 kV side)	+2% to -2%

Suitable loads :

- Resistive: Cookers, lights, heaters, kettles
- Motor: Fridges, washing machines, vacuum cleaners
- Electronic: TV, radio, florescent lights, PC's

10. FURTHER EXPERIENCE REQUIRED

At this stage there are still some uncertainties with this system as far as field experience is concerned.

10.1. LIGHTNING PERFORMANCE

The frequency of lightning surges which the system will be expected to survive may be very high. It is still too soon to say if the present design is robust enough from this point of view.

10.2. REAL CUSTOMER LOADING

The system needs to be used as a real power supply to real customers to establish the systems performance in an uncontrolled environment. This will be the next phase in the development of this type of power supply.

11. BENEFITS

11.1. ELECTRIFICATION

CAPTAP will play a significant role in bringing electricity to people in rural areas who live close to EHV Transmission lines, at a cost significantly less than other alternatives in many instances.

11.2. SOUTHERN AFRICAN GRID

CAPTAP is a means to give direct benefit to communities adjacent to AC transmission lines in neighbouring countries. This would have the effect of preserving the goodwill of people living near the line and thus curb vandalism and sabotage.

11.3. SERVITUDE ACQUISITION

In areas where new line servitude acquisition is difficult, CAPTAP may be used as a negotiating incentive.

11.4. TRANSMISSION LINE PERFORMANCE

CAPTAP is a means of transforming Extra High Voltages (EHV) like 400 kV and 765 kV down to distribution voltages like 22 kV and 11 kV for small loads. This is done without any physical connection to the transmission line phase conductors. Effectively, there is no danger to the performance of the main transmission line.

12. COST IMPLICATIONS

There are a number of factors which influence the cost of such a scheme. The prototype cost R137 000 in February 1993, but a significant price reduction can be expected once CAPTAP is commercially available.

13. CONCLUSION

Test results from the CAPTAP prototype near Apollo Substation are very encouraging. The technical performance of this induction supply is expected to be as good as a conventional single phase rural supply. At the end of the development phase, CAPTAP supplies will have definite application in the electricity supply market, the extent of which will be governed by its ultimate cost.

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THE EFFECT OF CHANGES IN AIR DENSITY ON CORONA FROM A SINGLE WATER DROPLET - MODELLING THE CORONA INCEPTION POINT

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This paper discusses the corona observed from a single water droplet suspended from a conductor at high potential. A model is presented which predicts the magnitude of applied electric field (E-field) necessary for corona inception as a function of air pressure. The model is used to show that two distinct processes result in a unique relationship between the applied inception E-field and pressure.

1. INTRODUCTION

Corona discharges from overhead transmission lines result in unwanted processes such as audio noise (AN), radio noise (RN) and corona losses (CL). In order to effectively design transmission systems it is important that the basic corona mechanisms are understood. In particular, the effects that water droplets suspended beneath a conductor have on corona processes should be studied.

The high permittivity of the water and the small radius of curvature at the tip of the droplet causes a significant enhancement of the E-field at the tip. Above a certain level of field enhancement, ionisation will result in the surrounding air. When this ionisation is self-sustaining, corona inception is considered to have occurred.

This self-sustaining discharge may occur in a manner similar to that which would occur at the tip of a conducting point. It is also possible that the droplet becomes unstable before the ionisation processes are dominant enough for a self-sustained discharge to occur. Under this condition, the small droplets that are torn from off the main droplet further enhance the field resulting in the reduction of corona inception.

Thus two distinctly different mechanisms result, and this produces a unique inception field versus air pressure curve. This curve is divided into two basic regions:

- the classical inception region (low values of air pressure)
- the instability region (higher values of air pressure)

Both these regions have been observed during measurement. Furthermore, the extension and tearing off of a smaller droplet from the main droplet has been observed at the higher pressures [1, 2, 3, 4].

2. THE MATHEMATICAL MODEL

The model allows a better understanding of these effects and facilitates the prediction of the behaviour of droplets suspended beneath a conductor. The model is briefly overviewed in this section. Further details may be found elsewhere [5, 6].

The model is used to determine the corona inception point of a droplet of known dimensions suspended in a uniform E-field. The model consists of three stages:

- determination of the shape of the droplet for a specific E-field
- determination of the magnitude of the E-field surrounding the droplet
- determination of the level of ionisation occurring as a result of the field divergence

Thereafter, a decision is made as to whether or not the inception of corona has occurred due to either instability, sufficient ionisation, or not at all.

2.1 THE SHAPE OF THE DROPLET

The droplet elongates in an E-field, and assumes a shape best modelled by an oblate or prolate spheroid. It is considered that the spheroid with the lowest potential energy for a specified E-field will best approximate the shape of the droplet under this condition [7].

An incompressible, inviscid half-spherical droplet of radius r_0 and permittivity ϵ_1 is suspended from a flat conductor in a medium of permittivity ϵ_0 and subjected to a uniform E-field, E_0 . The droplet elongates in the direction of the applied E-field into the shape of a half-spheroid.

The potential energy of the droplet is calculated by considering the sum of the four components of energy:

- surface tension energy (W_T)
- surface adhesion energy (W_{ad})
- gravitational energy (W_{grav})
- electrostatic potential energy (W_E)

The following equations were used in the determination of the total potential energy of the droplet:

The surface tension energy is given by [7]:

$$W_T = \frac{\pi a^2}{b^2} \sqrt{\frac{b^2}{a^2} - 1} \left[b \sqrt{\frac{b^4}{b^2 - a^2} - b^2} - \frac{b^4}{b^2 - a^2} \arcsin \left[\frac{b}{\sqrt{\frac{b^4}{b^2 - a^2}}} \right] \right] T \quad (1)$$

where: T is the surface tension [N/m]

a is the width of the droplet perpendicular to the E-field [m]

b is the length of the droplet, in the direction of the E-field [m]

Equation (1) is written in terms of a and b , but can be written in terms of b only as defined by equations (2):

$$a = \sqrt{\frac{r_o^3}{b}} \quad (2)$$

The surface adhesion energy can be evaluated in terms of the contact area of the droplet [8]:

$$W_{ad} = \pi a^2 T \cos \lambda \quad (3)$$

where: λ is the angle of contact between the liquid and the solid [rad]

The gravitational energy can be shown to be given by [9]:

$$W_{grav} = \frac{mg3b}{8} + C \quad (4)$$

where: m is the mass of the droplet [kg]

g is the gravitational acceleration constant [m/s²]

C is a constant defined by the initial height of the droplet above some reference

The electrostatic energy can be shown to be

given by [10, 11]:

$$W_E = \frac{1}{2} [\epsilon_o - \epsilon_i] E_i E_o \cdot \tau \quad (5)$$

where: E_i is the field within the droplet [V/m]

τ is the volume of the droplet [m³]

The field within the droplet may be calculated using Laplace's equation in both the oblate and prolate spheroidal plane, together with Legendre functions. This is more fully described elsewhere [5, 11].

The total potential energy is then determined by summing the energies given in equations (1), (3), (4) and (5). The total energy determined in this manner is plotted as a function of droplet elongation, b , in figure 1.

Furthermore, a minimum value for the total energy implies stability, while a maximum implies instability [7]. The minimum in the energy curve is evident from figure 1. From equation (2), the value of a corresponding to this value of b can be determined.

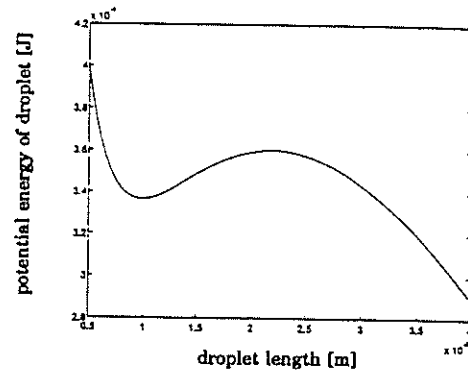


Figure 1 : Plot of the total potential energy of spheroids having the same volume but different lengths. The magnitude of the applied E-field remains constant.

2.2 THE EXTERNAL E-FIELD

The magnitude of the E-field external to the droplet, unlike the internal E-field, is not constant and can be calculated in both the oblate and prolate spheroidal planes using techniques described elsewhere [5, 11].

Using these techniques, the external field is readily determined. The field surrounding a prolate spheroidal droplet with a relative permittivity of 80 is shown in figure 2. The E-field is always a maximum at the tip of the droplet, and the magnitude of the field along a straight line away from the tip is shown in figure 3. This data allows determination of the level of ionisation.

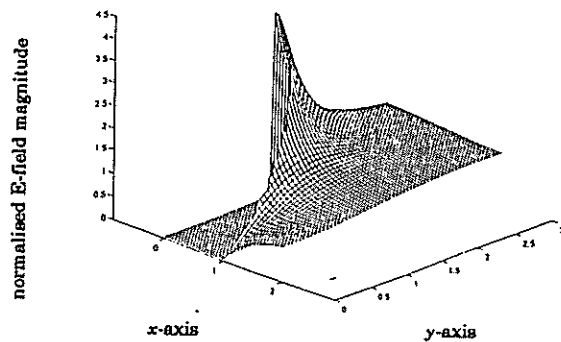


Figure 2 : A 3-D plot of the normalised magnitude of the E-field external to the droplet. The droplet is 1,5mm in length, and 1mm in width.

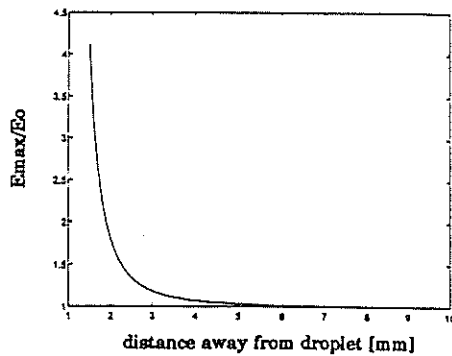


Figure 3 : Plot of the normalised magnitude of the E-field as a function of distance in a straight line away from the tip of the droplet.

2.3 THE CORONA INCEPTION FIELD

The value of applied E-field at which inception will occur is determined by calculating the number of electrons along a straight line away from the droplet produced as a result of ionisation due to a single primary electron. When the number of electrons at any point along this lines is found to exceed $1,8 \times 10^3$, inception is considered to have occurred [12, 13].

The attachment and ionisation coefficients for air are known as functions of pressure. Furthermore, the relationship between the droplet shape and applied field magnitude, as well as the magnitude of the field in a straight line away from the droplet are known. Hence the number of free electrons can also be determined as a function of the distance from the droplet for a specific value of pressure.

It must also be noted that the integral of $(\alpha - \eta)$ will reach a maximum at some point d from the cathode. This is evident from figure 4, which shows that $(\alpha - \eta)$ becomes negative beyond the point d . From this it can be determined whether or not corona inception has occurred. The procedure is repeated for varying values of air pressure, allowing construction of the curve shown in figure 5.

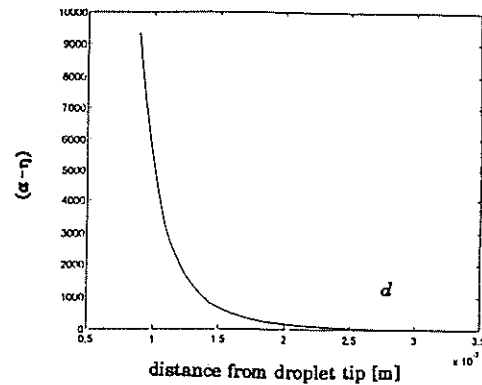


Figure 4 : Plot of $(\alpha - \eta)$ as a function of distance in a straight line away from the droplet.

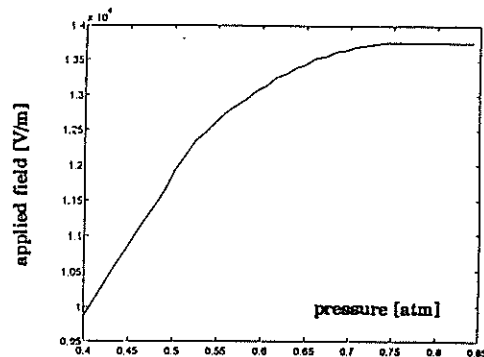


Figure 5 : Plot of the magnitude of the applied E-field necessary to result in the inception of corona from a water droplet of radius 1,25mm as a function of air pressure.

3. COMPARING MEASURED AND PREDICTED RESULTS

The inception of corona from a single droplet in both a divergent and a quasi-uniform field has been recorded as a function of air pressure. The pressure range over which these measurements were taken was 0,4 - 0,8 bar. These results, as well a details of the experimental set-up, are fully described elsewhere [3, 4].

Figure 6 shows an example of the measured inception point. This curve should be compared with that shown in figure 5. It is evident that the inception point does not change significantly with pressure until pressures below 0,8 bar are reached. This trend is similar to that shown in figure 5. While it can be seen that the knee of the curve for both the measured and the predicted case occurs between 0,6 and 0,7 atmospheres, the magnitude of the applied field is greater in the model than was actually recorded. This can be ascribed to the divergency of the local field at the droplet tip in the experimental case.

4. DISCUSSION OF RESULTS

It is evident from both the measured and predicted results that for higher values of air

pressure the magnitude of the E-field necessary for corona inception does not vary significantly. At lower pressures, this is not the case.

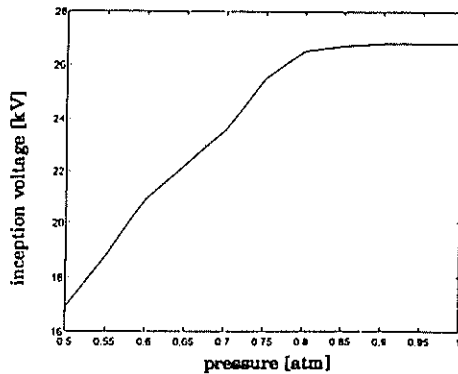


Figure 6 : Plot of the measured inception point of corona from a water droplet on a negative DC electrode as a function of air pressure.

This can be explained by noting that two different mechanisms are responsible for corona inception:

- inception due to sustained ionisation of the surrounding gas with no instability of the droplet
- inception as a result of droplet instability

The development of the mathematical model, however, has allowed a better appreciation of these two regions, and has further allowed a third region to be defined.

At the higher pressures, the droplet becomes unstable before sufficient free electrons for sustained ionisation exist. The droplets torn from the main droplet cause further field enhancement and hence ionisation.

At the lower pressures, sufficient free electrons for sustained ionisation exist prior to the conditions for droplet instability being met. However, this region can be further subdivided into two distinct regimes. At the low values of E-field shown in figure 5 the inception field changes almost linearly with air pressure. This is because the effect of droplet elongation on the inception process is not predominant.

At higher values of field, the elongation begins to dominate, and the linear relationship no longer holds.

8. CONCLUSION

The model presented in this paper has shown that the inception of corona from water droplets suspended beneath a conductor does not follow the trends associated with a conducting electrode. The elongation and instability of the droplet has been shown to have a marked effect.

A good correlation between measured

and predicted results has been found. This establishes the validity of the modelling method presented here.

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LIVE-LINE WORKING ON OVERHEAD POWER LINES: AN ASSESSMENT OF THE NEED FOR FURTHER RESEARCH

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ABSTRACT

The working practices for the live-line maintenance, repair and replacement of the components of high-voltage lines and substations are briefly reviewed; particularly with regard to safety features. The aspects requiring further investigations have been identified as a prelude to initiating a research programme.

1. INTRODUCTION

With the ever increasing cost of down time on overhead lines, methods are being developed to effect repairs and maintenance by using live-line techniques [1]. Obviously, safety is a very important aspect [2].

Live-line maintenance can be separated into two broad categories:

- * Washing of insulators (preventative maintenance)
- * Repairing and replacing of hardware (e.g. insulators)

In Natal, insulator washing has become the accepted method of preventing faults due to pollution on the insulators caused by cane fires, salt mist etc [3].

There are two acceptable techniques of washing of the insulators in situ using deionized water. The first is airborne spraywashing from a helicopter. The other is ground-based washing for substations, done with the affected bays isolated. Hand-held equipment or a fixed installation is usually used.

For the replacement of insulators, two practices are employed. With the "hot-stick" method, an insulating rod is used to hold the various tools required for the operation. The other practice is the

"bare hands" technique whereby the worker is brought to line potential.

2. LIVE-LINE WASHING

The four general methods in use for hot line washing differ mainly by the type of nozzle arrangement [4]:

- * Portable hand-held jet nozzle
- * Remote controlled jet nozzle
- * Fixed spray nozzles
- * Helicopter mounted nozzles

The efficacy and safety of the washing technique is influenced by nozzle-conductor distance, quantity of wash water, effect of wind, type of nozzle and washing process. There is scope for undertaking further investigations into:

- (a) The washing process.
- (b) Dielectric strength with overvoltages.
- (c) Simulation with actual contaminants, rather than standard pollution tests.
- (d) Washing under DC energization.

3. LIVE-LINE MAINTENANCE

The hot-stick method of live-line maintenance has been used since about 1913 when "disconnect sticks" were employed for opening energized isolators on lines [5]. These hot-sticks were originally made of wood but have now been superseded by ones made of resin bonded glass fibre, which provide superior performance in respect to permeability of moisture and mechanical strength. These newer hot-sticks, however, need further development to achieve a good stiffness combined with lightness.

With the bare-hands method, safety is of paramount importance as any flashover occurring to or near the worker could be

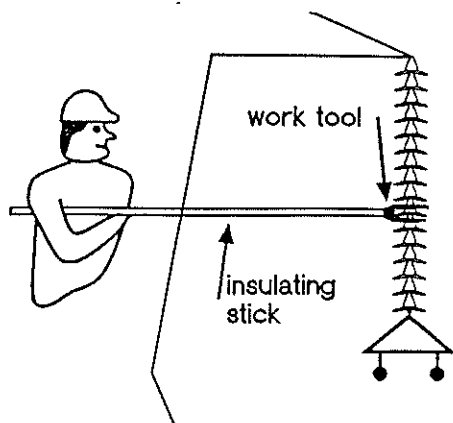


Fig 1: Live-line working with man at earth potential.

fatal. This safety is ensured by adherence to strict training practices rather than making the line safe by means of earth straps.

4. LIVE-LINE CLEARANCE

For a person working on a live line (Fig. 1), the proper electrical clearances obviously need to be maintained. If they are too large, dexterity with the equipment is lost; if they are too small, a harmful flashover may occur. The difficulty lies with specifying the minimum clearance for an acceptable flashover-probability. That is, the "statistical withstand" voltage for the minimum air-insulation distance must be known.

To generalise the data, the usual practice is to relate – by a gap factor, k – the 50% flashover voltage, U_{50} , of any gap geometry to that of a rod-plane gap for the same distance [6]; i.e.

$$k = \frac{U_{50}}{U_{50(\text{rod-plane})}} \quad (1)$$

The value of U_{50} (rod-plane) for a certain distance is a function of both the type of energization and the ambient conditions (i.e. air density and relative humidity) whereas k depends only on the gap geometry. Therefore, for the various working arrangements, the corresponding value of k needs to be determined. By this approach, the large amount of data existing for the rod-plane gap can, hopefully, be used for other geometries.

To date, the gap size corresponding to a low flashover probability (e.g. 0.15%) has been calculated by using Gaussian statistics at the $U_{50}-3\sigma$ value – where σ is the standard deviation [7]. The corresponding "risk of failure" for a given overvoltage function is then established by using the standard method [6]. It seems unlikely, however, that the probability function of the flashover voltage will be Gaussian in the region of relevance to live-line working because electric strength has a minimum-value property. That is, the probability function is more likely to be negatively skewed and so Extreme-value statistics [8] rather than Gaussian statistics should be used. Although this aspect is rarely pursued for the "run of the mill" insulation co-ordination cases, the consequence of having a flashover during live-line working is so serious that it warrants careful consideration. Although some research has investigated this low-probability region [9], much more information is required to establish a definitive viewpoint.

For such rare events, the flashover probability $p(d)$ of a rod-plane gap of distance d at a voltage U is possibly of the form:

$$p(d) = C \left(\frac{U - U_0}{U_0} \right)^n \quad (2)$$

where C , n and U_0 are constants that need to be determined by experimentation. Although such an exercise will be highly time consuming, automated control and measurement would make it substantially less laborious.

Of particular concern for live-line workers in South Africa, is the influence of reduced air density (eg. ~ 17% reduction on the Reef) and of high relative humidity (eg. near saturation levels in Natal Province during summer).

5. FLOATING OBJECTS

Taudiere and Lalot [10] have investigated, for switching surges, the consequence of having metallic bodies at floating potential in the air gap between two

phases. The breakdown voltage of the split gap – i.e. the combined air gap on both sides of the floating body – is substantially lower (up to 30%) than that for a single gap of the same length. They have also shown that the object's position in the gap has a bearing on the flashover voltage, with this being the lowest when the body is situated close to the positive side. These findings lead to the questions of what happens when: (a) bodies of different conductivities are inserted into the air gap? and (b) different types of impulses are applied?

6. DEFECTIVE INSULATORS

A complicating feature is the reduction in the flashover strength of a string of disc insulators due to it containing some faulty units. For glass insulators, the shattered units are immediately apparent but punctures or cracks in porcelain insulators can usually not be detected by a cursory visual inspection – although other techniques are available [11].

Some investigations [2] [12] have established the importance of the number of broken insulators in the string, the type of insulator–damage and the position of the damaged units within the string. Although other research into this topic is known to be taking place [13], it is so important that additional effort is probably required.

Therefore, for safe working, the determination of the statistical withstand voltage must take into account defective insulator strings.

7. PROTECTIVE DEVICE

To provide additional safeguards, protective devices can be temporarily fitted on either side of the work-point at positions sufficiently remote to avoid their operation, endangering the live-line workers. That is, an overvoltage wave that propagates along the line should be reduced appreciably when it reaches the work-point. However, such a device must not cause nuisance tripping of the circuit breakers due to minor transients.

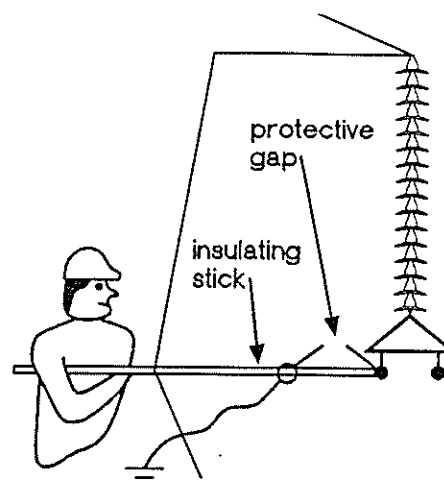


Fig 2: Protective gap being temporarily attached to a line.

The crux of the design and/or setting of the protective device is that it has to operate reliably for these opposing requirements.

To date, the practice is to use simple air gaps placed on towers (Fig 2) adjacent to the span in which work is underway. van Name [7], proposes that the statistical withstand voltage for the minimum air–insulation distance must be greater than the 2σ (i.e. 97.7% probability) flashover voltage of the protective gap. The consequence of using this value or of increasing the safety margin to the more usually specified 3σ value (i.e. 99.85% flashover probability) for insulation co-ordination [6] requires to be examined once the low-probability aspect has been thoroughly investigated.

Two alternative approaches that warrant consideration for improving the safety of such a working arrangement are to replace the simple air gap with either a toroidal gap or a surge arrester. The former has been developed in the UK for the improvement of insulation co-ordination [14]. The "gapless" zinc oxide arrester has good potential for reliability and accurate control; its further advantage is that it does not short the line to earth but rather clamps the voltage at the desired level whilst shunting off the excess charge. Although such an arrester is available for voltages up to at least 400 kV, the standard design with the porcelain housing is heavy.

However, a much lighter version, having a polymeric housing, exists for use up to about 66 kV [15]. It seems possible that such a light-weight device could be developed for the higher voltages.

8. CONCLUSIONS

Although much has been accomplished to ensure that live-line working on high-voltage equipment is both profitable and safe, quite a few aspects need to be investigated further. The most important ones involve the prevention of flashover for lightning waves and switching surges. Although live-line working does not take place when lightning is in the vicinity, experience from abroad shows that very remote strikes can result in impulse overvoltages being propagated over long distances. Therefore, research needs to be undertaken to determine:

- (a) More accurate values for the gap factor for relevant geometrics.
- (b) The low-probability region of the flashover-voltage function.
- (c) The reduction in the flashover voltage caused by conducting and insulating objects at floating potential in the air gap.
- (d) The influence of reduced air density and high humidity.
- (e) The electrical degradation of insulator strings due to some defective units.
- (f) The design of a superior protective device; e.g. a toroidal co-ordinating gap, a light-weight zinc oxide surge arrester.

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Calibration of a Capacitive Voltage Divider for the Measurement of Very Fast Transients (VFT's) in Bus Ducts.

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Abstract

The high frequency, low frequency and transient response of a split capacitance voltage divider, used to measure VFT's, was obtained to determine the effect the divider had on the accuracy of experimentally measured VFT V-t curves. It was found that the divider used had a mismatch between the coaxial cable of the divider and the rest of the low voltage arm of the divider. This mismatch slowed the response of the divider, when subjected to a fast rising voltage step, and resulted in inaccurate measurement of the experimentally generated VFT V-t curves. Only once all the uncertainties in the measurement of the VFT waveforms have been removed can any meaningful conclusions be drawn about the relationship between breakdown and voltage waveform.

1. Introduction

There have been reports of failures in GIS (Gas Insulated Substations) associated with the operation of disconnect switches /1,2/. Some researchers have reported that VFT (Very Fast Transient) initiated failures in a commercial-grade system can take place at a voltage well below the lightning impulse withstand level of the system /3/. Other researchers have found no difference between the V-t (volt time) curves measured for lightning impulses and fast rising impulse voltages /4/. A VFT, however, has a steep front and a distinct oscillatory nature, and it has been proposed that there may be a relationship between the breakdown voltage and the frequency content of the VFT /5/.

During disconnect switch striking, the voltage over the disconnecter gap collapses within 2-12ns. The speed of the voltage collapse depends on the disconnect switch design, gap distance and the gas pressure in the disconnect switch. This voltage collapse gives rise to two travelling voltage surges of opposite polarity, moving away from both sides of the disconnect switch. There will be reflections of these travelling waves from all points of impedance discontinuity within the GIS system. These surges and their associated reflections interact with one another to form the complex disconnect switch generated VFT in the system /6,7,8/. Since the VFT is comprised of the initial voltage surges and all the reflections of these initial surges from all points

of impedance discontinuity within the system, the characteristics of the VFT will depend on the geometry of the system.

As can be seen from the above discussion, a VFT must be treated as a voltage travelling wave with a wide frequency range. This must be borne in mind when measuring its waveform.

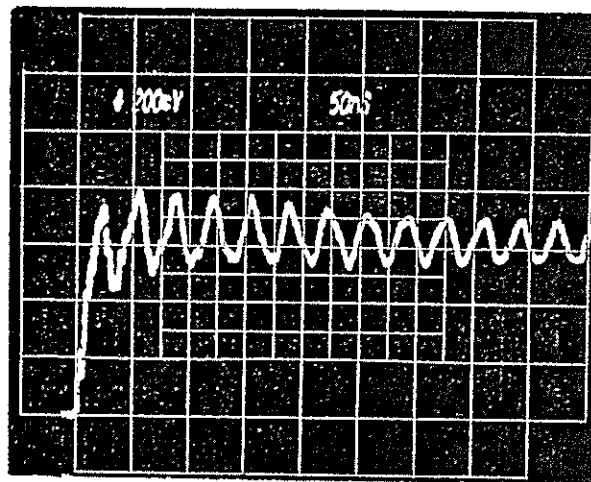


Figure 1. Experimentally generated VFT. Scale 400mV per division. Time base 100 ns per division.

In order to accurately establish the VFT V-t curves for any gap configuration, the voltage waveform in the gap must be accurately known. Due to practical constraints it is not always possible to measure the voltage directly in the gap under consideration /9/. This results in the voltage divider being remotely located at some other point in the experimental test vessel. Due to the travelling wave nature of a VFT, a correction will then be necessary to the remotely measured voltage to relate it to the voltage in the test gap. This correction is the topic of ongoing research.

Figure 1 shows a typical experimentally measured VFT voltage waveform. This was measured at the remote location of the capacitive voltage divider, and was generated in the experimental test vessel, described elsewhere, with no breakdown in the test gap /9/.

The method used to construct the experimental VFT V-t

curves is that used by Reynders /3/. This method uses the highest voltage attained by the VFT before or at breakdown. From the experimental investigation it was determined that the fastest time to breakdown obtained was 120ns with a SF₆ gas pressure of 2 bar absolute in the 180/110mm coaxial test gap. From *figure 1* it appears that the second peak of the oscillation of the VFT is the highest voltage reached before breakdown and should therefore be taken as the breakdown voltage. It will therefore be necessary to accurately know the response of the capacitive voltage divider during the first 200ns in order to determine if the second peak of the measured VFT is in fact the highest voltage reached before breakdown.

This paper is concerned with establishing the accuracy of the capacitive voltage divider used to remotely measure the experimentally generated VFT voltage waveform. Only when all areas of uncertainty in the measurement of the VFT waveform have been accounted for can any meaningful conclusions be drawn about the effect of the VFT wave shape on the breakdown voltage.

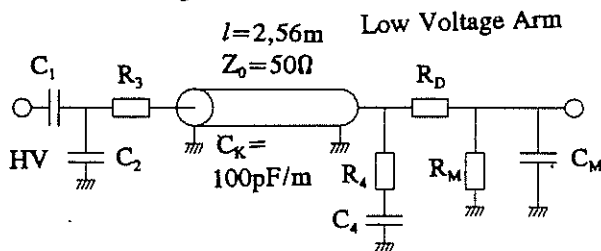


Figure 2. Schematic of capacitive voltage divider.

2 Capacitive Voltage Divider

Figure 2 shows a schematic of the capacitive voltage divider used to measure the VFT in *figure 1*. This is a split-capacitance voltage divider. The fundamental operation of this type of capacitive voltage divider can be found elsewhere /10, 11/. R_M and C_M are the resistance and the capacitance presented to the divider by the measurement oscilloscope.

3 High Frequency Response

The high frequency response of the capacitive voltage divider was determined by subjecting the divider to a fast rising voltage step.

3.1 High Voltage Pulse Generator

The requirements of the voltage step used to determine the step response of the capacitive voltage divider are as follows:

1. The step must have sufficient amplitude such that the output of the divider can be easily extracted, with confidence, from the noise in the system. Due to the 227 000:1 nominal

division ratio of the divider it was necessary to use a voltage step with a amplitude larger than 1,5kV. A input voltage of 1,5kV will give an output from the divider of 6,61mV at a nominal division ratio of 227 000:1. This is of sufficient amplitude, above the noise level, to allow for the accurate extraction of the high frequency response.

2. The rise time of the step must be equal to or faster than the rise time of the fastest event that will be measured by the divider. This rise time must however not be fast enough to excite the test set up into modes of propagation other than the TEM mode, as this will give erroneous results. From *figure 1* it can be seen that the first step of the VFT has a rise time of approximately 5ns. From these considerations the voltage step had a rise time of approximately 4ns.
3. The top of the step must be a flat as possible and as free from reflections as possible during the time of interest.

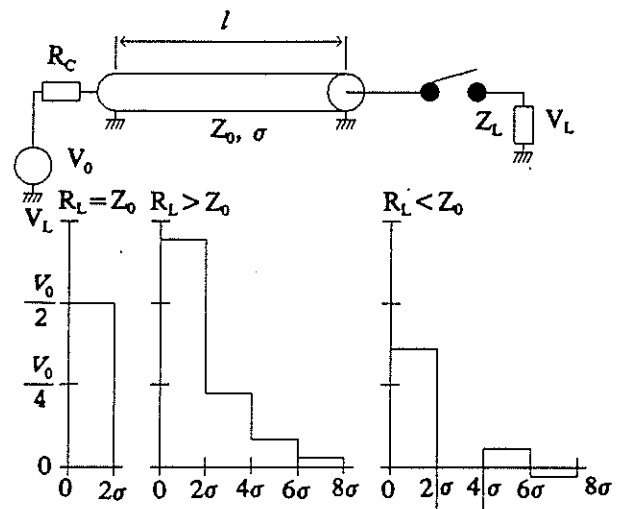


Figure 3. Blumlein pulse generator.

Where:

V_0 = Charging voltage.

R_C = Charging resistor of approximately 10MΩ.

Z_0 = Transmission line characteristic impedance.

σ = Transmission time of the transmission line of length l .

Z_L = Impedance of the load.

V_L = Voltage across load Z_L .

It was decided to use a Blumlein type pulse generator as it could satisfy the above needs. *Figure 3* shows the basic principal behind the generator. The transmission line is charged up to a voltage of V_0 . This line has a characteristic impedance of Z_0 and a transmission time of σ . The load has a impedance of Z_L . The switch must be fast and free of contact bounce. The output voltage waveforms for different characteristic load impedances can be seen in *figure 3*.

3.2 Experimental Set Up

The capacitive voltage divider must be calibrated in the same geometrical set up as that in which it is to be used. This particular divider is fitted in a port in a disconnect switch enclosure. Figure 4 shows the experimental set up used to determine the step response of the divider. Impedance matching cones, with a constant ratio of radii, were used to match the generator to the capacitive voltage divider. The 200m of coaxial cable at the end of the divider is used to prevent reflections of the generated step wave during the time of interest.

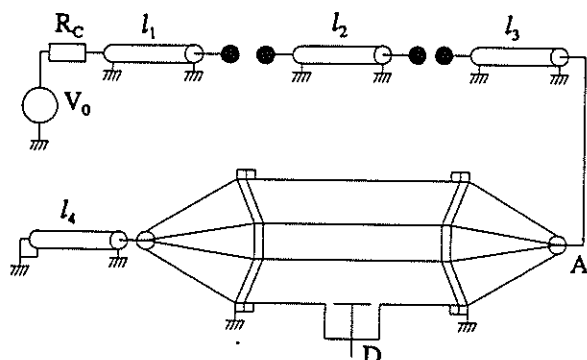


Figure 4. Step response experimental set up.

Where:

V_0 = Charging voltage.

R_c = Charging resistor of $10M\Omega$.

l_1 = 100m of coaxial cable.

l_2 = 10m of low loss coaxial cable.

l_3 = 2m of low loss coaxial cable.

A = Input voltage measurement point.

D = Voltage divider under test.

l_4 = 200m of coaxial cable.

Figure 5 shows the generated voltage step measured at point A in figure 4. This voltage step has a rise time from 10% to 90% of 4ns and an amplitude of 1,69kV. The duration of the step is $1\mu s$. The voltage step in figure 5 was measured using a 4kV voltage probe with a bandwidth of 400MHz and a rise time of 0.8ns.

3.3 Results

Figure 6 shows the output of the capacitive voltage divider, the input voltage step is shown in figure 5. From figure 6 it can be clearly seen that there is an impedance mismatch between the coaxial cable of the divider and the rest of the low voltage arm of the divider, which has resulted in the clearly visible reflections. Until the cable has been correctly matched to the low voltage arm of the divider no meaningful high frequency division ratio can be obtained.

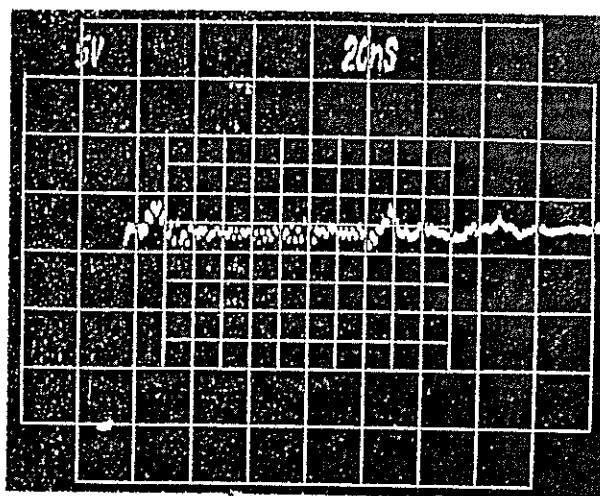


Figure 5. Generated voltage step. Scale 500V per division. Time base 20ns per division.

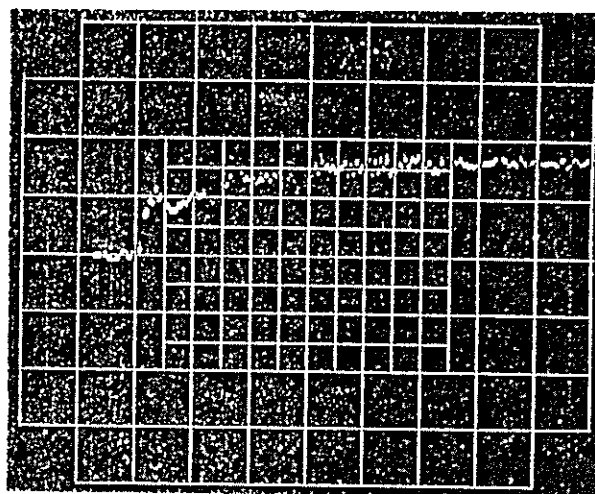


Figure 6. Output of divider for the input voltage in figure 5. Scale 5mV per division. Time base 20ns per division.

4 Low Frequency Response

The low frequency response of the capacitive voltage divider was determined under 50Hz conditions in the same geometrical set up as that in figure 4. From this the nominal division ratio of the divider was determined to be 227 000:1.

5 Discussion

From the results of the high and the low frequency responses of the capacitive divider it was decided to rebuild the low voltage arm of the divider using new components. This resulted in the elimination of the reflections that were present in the step response of the divider using the original low voltage arm. The high and low frequency division ratios coincided at 227 000:1 for

the reconstructed divider.

The reflections present in the original capacitive divider result in a slow rise of the output voltage from the divider for the first 100ns under transient conditions. From *figure 1* it can be seen that the first peak of the VFT occurs in the first 50ns. This leads to the conclusion that the first peak of the VFT might be the highest voltage reached by the VFT due to this peak being attenuated in size by the reflections due to the mismatch. This conclusion must still be verified by further experimental work.

It has been shown that there is an error introduced into the measured VFT V-t curves due to the divider not being located at the point of break down [9]. This error was determined by implementing a computer model of the test vessel using ATP and comparing the simulated VFT waveform in the center of the test gap with that at the location of the divider. For this particular test vessel geometry it was found that the experimentally measured VFT V-t curves must be corrected up in amplitude by approximately 11% if the amplitudes of the second peaks of the VFT waveforms are compared. If it is found that the first peak of the experimentally generated VFT waveform is in fact the highest voltage reached before breakdown, a correction can then be found by comparing the amplitudes of the first peaks of the simulated waveforms.

6 Conclusions

In order to be able to produce accurate VFT V-t curves the experimentally generated VFT voltage waveform must be accurately measured. To be able to do this it is essential to know that the high and the low frequency division ratios of the divider, over the frequency range of interest, are equal. It is also essential to determine the transient response of the divider so that it can be determined that the cable of the divider is correctly matched to the low voltage arm of the split capacitance divider.

By applying a voltage step to the system with a magnitude of 1,69kV and a rise time from 10% to 90% of 4ns it was determined that there was an impedance mismatch between the low voltage arm of the divider and the coaxial cable of the divider. This resulted in the reconstruction of the low voltage arm of the divider using new components.

These reflections present in the output of the divider, due to a step voltage, result in the output voltage rising slowly. This could give rise to the experimentally generated VFT voltage waveforms having a higher first peak rather than a higher second peak. This will change the magnitude of the breakdown voltage used to construct the VFT V-t curves.

From this work it has been clearly shown that there is

an attenuation of the measured voltage waveform over the time period of interest. This can introduce significant error into the experimentally measured VFT V-t curves. Only once all the uncertainties in the measurement of the VFT waveforms have been removed can any meaningful conclusions be drawn about the relationship between breakdown and voltage waveform.

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DEVELOPMENT OF A COMBINATION GENERATOR FOR LIGHTNING SIMULATION

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This paper discusses the development of a combination generator for the generation of pseudo-lightning surges. The generator is capable of producing an 8/20 μ s current impulse with a 15 kA peak, and a 1,2/50 μ s voltage impulse with a 15 kV peak. Emphasis is placed on a description of the design and manufacture of the various system components, such as the high speed high current measuring circuit. A novel method of winding suitable low-inductance high power resistors is presented.

1. INTRODUCTION

The study of the effect of lightning on industrial systems is gaining in prominence, and has been the subject of extensive research [1, 2, 3]. In Southern Africa, a combination of industrial development and increased installation of electronics into industrial sites has contributed to a substantial increase in the awareness of the problem.

The adequate protection of industrial sites against the effects of lightning activity encompasses a number of factors:

- correct system earthing
- correct cabling strategies within buildings
- correct cabling strategies between buildings
- the effective zoning of equipment in terms of risk of failure
- incorporation of adequate surge suppression devices into the system

The first three aspects have traditionally been emphasised in National Standards and standard texts on noise suppression [4, 5, 6], as well as in manufacturers' installation documents. Furthermore, these aspects are being incorporated into comprehensive International Standards [7].

However, a very important aspect involves the specification of the correct surge suppression device. To this end, it should be appreciated that while the energy handling capability of such a device can be calculated for various earthing, bonding and cabling strategies [1], the testing of such devices remains a significant challenge. New standards, however, should assist in clarifying this problem [7].

A number of test circuits have been proposed, of which the standard 8/20 μ s current waveform and the 1,2/50 μ s voltage waveform have been the most common.

A combination generator produces a surge whose nature is dependent upon the impedance of the load. A major motivating factor for using this generator is that it produces a significantly more realistic pseudo-lightning surge than that produced by a dedicated voltage or current generator, allowing more realistic testing of surge suppression devices [8].

2. LIGHTNING PROTECTION

It is important to note that any installation must include both internal and external components of lightning protection. Whereas the external protection is well documented in National and International Standards [4, 8], internal protection is often not considered.

It should be stressed that the fundamental objective of internal lightning protection - or indeed

protection of systems running between two buildings - relies for its effectiveness upon the establishment of an equipotential platform [3, 9].

The concept of zoning of equipment in terms of the harshness of the lightning effects it is likely to encounter blends closely with this concept. This implies, however, that the boundaries between zones, as well as the equipotentialization, must be achieved by means of bonding conductors or surge suppressors connecting the equipment earth to the building earth at the point of entry to the boundary.

This has necessitated the testing of surge suppressors. While a number of waveforms and generator types have been proposed [7, 8], the combination generator described here most effectively produces that type of response in a suppressor that would be effected by a lightning surge.

3. DESCRIPTION OF THE CIRCUIT

The ideal, theoretical, simplified circuit [8] is shown in figure 1.

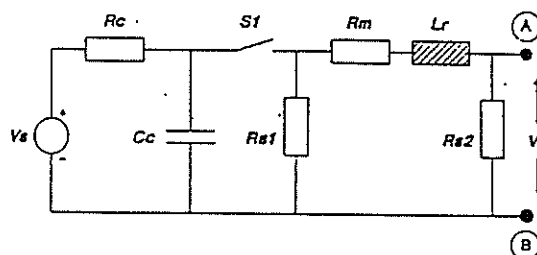


Figure 1: The ideal, theoretical, simplified circuit of the combination generator

Via the resistor R_c , the voltage source V_s charges the capacitor C_c which discharges into the remainder of the circuit when the spark gap S_1 closes. R_s is the pulse duration shaping resistor, R_m is the impedance matching resistor, and L_r is the rise-time shaping inductor.

The open-circuit output voltage V_o is developed across R_{s2} which also affects the rise-time of the voltage waveform; an increase in the value of R_{s2} will result in a shorter rise-time of the voltage waveform. R_{s1} affects the fall-time of the voltage waveform; an increase in its value will lengthen the falling edge of the voltage impulse. For a functional circuit, the value of R_{s1} is much greater than that of R_m . This results in R_{s1} having a negligible effect on the current waveform, and R_m having a negligible effect on the voltage waveform. In fact, the functional short-circuit equivalent circuit of the combination generator consists only of C_c , R_m and L_r (R_{s2} is short-circuited).

The high currents and voltages that R_{s2} has to withstand make it imperative to use a high power, wirewound resistor for this component. Consequently, a significant series inductance is inevitable, as is a small amount of distributed capacitance. For a

practical model, the capacitance is small enough to be neglected, but the series inductance has to be accounted for. Therefore, the most functional equivalent circuit (figure 2) has high value resistors (greater than $5\ \Omega$) represented by a pure resistor and pure inductor in series. This circuit allows simple simulation yet it produces results that are sufficiently accurate for the purposes of design.

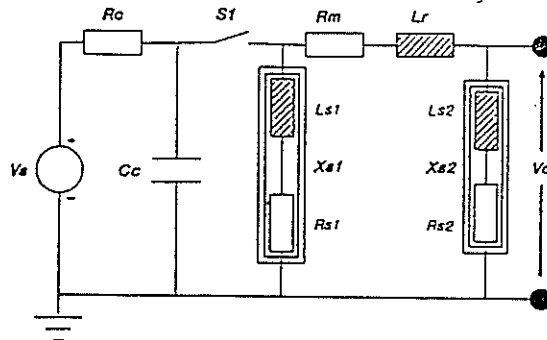


Figure 2: Practical, simplified equivalent circuit of the combination generator

3.1 COMPONENT SELECTION

It would be time-consuming and unnecessary to produce a computer program that could automatically determine the required component values of the generator. An understanding of the functions of each component, together with a simple numerical simulation program, allowed suitable values to be selected with relatively few iterations.

The program which was produced for the purpose is based on the extraction of the time-response of the circuit from the state equations, using Euler's method. The input data consists of the parameters in figure 2; the output is a graphical representation of the voltage and current waveforms.

The most effective design procedure was to select a value for the capacitor C_c , and to begin with the current impulse equivalent circuit, yielding values for R_m , L_r , and obviously C_c (R_{s1} is neglected for the current impulse circuit).

Subsequently, considering the voltage impulse in particular (the complete equivalent circuit), the values of X_{s1} and X_{s2} were determined. Since X_{s2} has a significant effect on both the current and voltage waveforms, the objective of the design involved repeated iteration until a suitable value of X_{s2} was determined. It was noted that, even for relatively large values (exceeding $50\ \mu\text{H}$), L_{s1} had a negligible effect on both the voltage and current waveforms. It was therefore considered unnecessary to design for a particular value of L_{s1} , provided that it was less than $50\ \mu\text{H}$.

The component values that provided the most effective compromise are illustrated in figure 3.

3.2 THE CURRENT MEASUREMENT CIRCUIT

Since one output of the combination generator is earthed, galvanic isolation is not imperative. However, to protect associated equipment, an isolated measurement circuit is a sensible option. A resistive shunt is therefore unsuitable, and a standard current transformer will not have sufficient bandwidth. Additionally, commercially-available hall probes with

sufficient bandwidth do not have the required current rating.

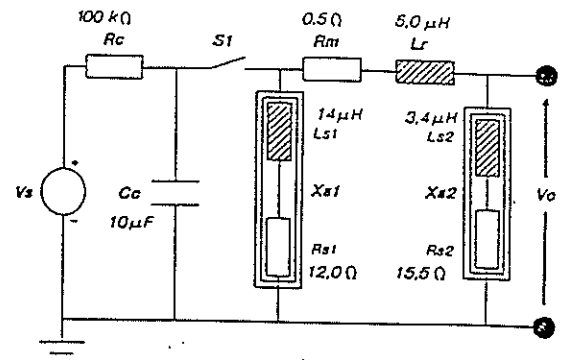


Figure 3: The selected component values of the combination generator

The only available alternative was to use a Rogowski coil coupled to a passive RC integrator, as shown in figure 4. The coil itself consists of 700 turns wound on a 25 mm diameter plastic tube, with one end of the coil routed to the other through the centre of the winding, in order to reduce stray effects.

The RC integrator has a time-constant of approximately $900\ \mu\text{s}$, resulting in a bandwidth extending down to a few kilohertz. The Rogowski coil compares very favourably with a *Pearson Type 301-X high speed pulse current transformer*, except for a small amount of spark gap-induced interference at the beginning of the impulse (the first few μs).

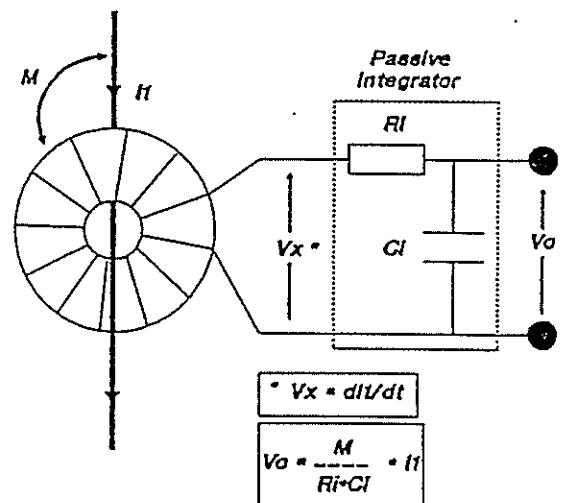


Figure 4: The Rogowski coil current measurement circuit

4. FABRICATING THE COMPONENTS

The resistors were wound using 0.9 mm *Nichrome* resistance wire with a resistance of $1.68\ \Omega/\text{m}$, insulated with glass/polyester sleeving with an insulation rating of 4 kV and a maximum temperature of 200°C . Flat, rectangular *perspex* formers were used

to minimise series inductance by virtue of their small cross-sectional areas.

The capacitor C_c consisted of three Westingcorp 30 μF , 10 kV impulse capacitors connected in series.

The two important challenges were the design of the high power, low-inductance resistor R_{s2} , and the automatically triggered spark gap switch.

4.1 THE LOW-INDUCTANCE RESISTOR

A difficult aspect of the design was to fabricate the high voltage, high power resistor R_{s2} with sufficiently low series inductance.

Experimentation showed that both the Zig-zag and the Ayrton-Perry [10] methods (figure 5) did not provide sufficiently low series inductance. While a bifilar (figure 5) wound resistor provided low enough series inductance, it failed to withstand the high voltage impulse using the existing 4 kV sleeving, and thicker sleeving simply increased the inductance.

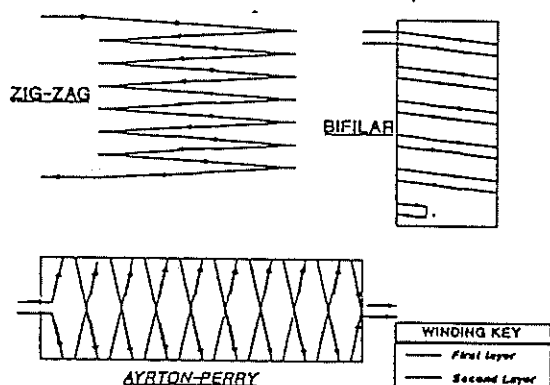


Figure 5: Existing low-inductance winding techniques for wirewound resistors

After much experimentation with various winding arrangements which allowed current in adjacent turns to flow in opposite directions, a novel method was realised which had less than half the series inductance of a similar resistor that was wound using the existing methods. The method chosen is shown in figure 6.

The prototype resistor R_{s2} was wound with a measured series inductance of 3,4 μH . Since this was the lowest value obtainable, it was therefore not a variable parameter in the design process, so L_{s2} (figure 3) necessarily assumed this value when the component values were determined.

4.2 THE SPARK GAP SWITCH S_1

An important aspect of the design was to fabricate an adjustable, automatically triggered spark gap capable of conducting the high currents without damage.

The selected method involved flame-spraying copper electrodes with tungsten-carbide. This gap has conducted in excess of 5000 10 kA current impulses with negligible signs of physical degradation.

The triggering system is shown in figure 7. The spark gap electrodes are mounted on a pivoted arm which is actuated by a 220 V solenoid. The solenoid is controlled by the comparator and sensitivity adjustment circuit. The circuit includes several low-pass filters to avoid noise-induced triggering of the gap.

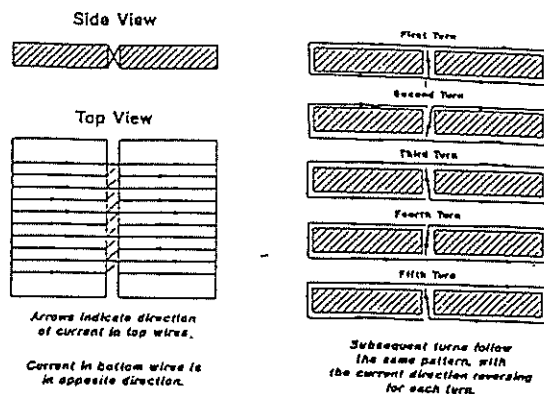


Figure 6: The novel method of winding the low-inductance resistors

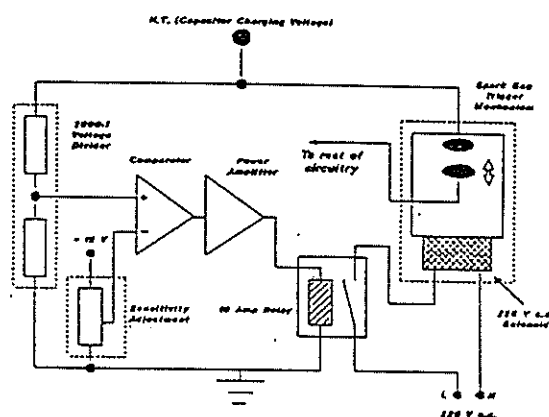


Figure 7: The spark gap control system block diagram

5. DISCUSSION OF RESULTS

5.1 THE VOLTAGE WAVEFORM

The predicted voltage waveform is compared with the measured waveform in figure 8.

VOLTAGE WAVEFORMS Ideal Versus Obtained Waveforms

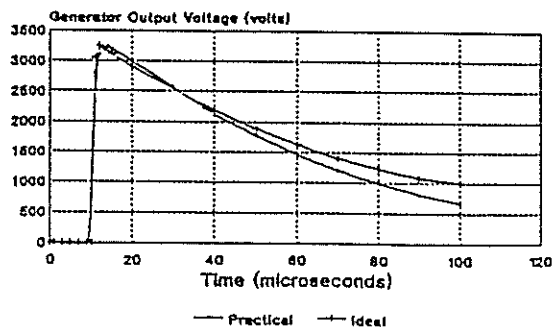


Figure 8: The predicted and measured voltage waveforms

The voltage waveform varies slightly from impulse to impulse but, in most cases, it is within the

permitted specification [11]. The mean waveform is a $1.55/45 \mu\text{s}$ impulse, with some transient oscillations at the leading edge of the waveform; these were reduced considerably when the earthing system was rearranged.

It is important to note that the optimum (ideal) $1.2/50 \mu\text{s}$ impulse with no leading edge oscillations waveform will only be possible if R_{s2} (figure 3) can be fabricated with negligible series inductance or distributed capacitance.

5.2 THE CURRENT WAVEFORM

The predicted current waveform is compared with the measured waveform in figure 9.

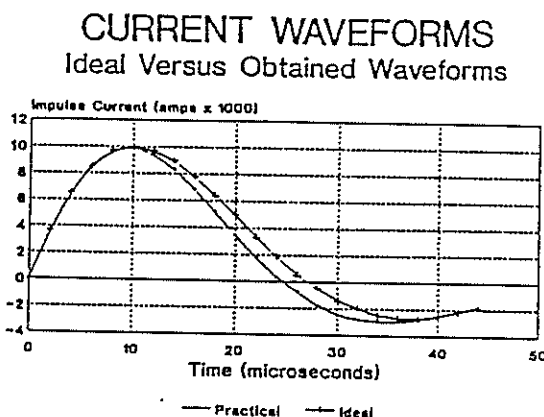


Figure 9: The predicted and measured current waveforms

The current waveform varies slightly with different current levels but, again, most of the impulses are within the permitted specification [11]. The deviation from the ideal waveform is due entirely to the compromise between producing a good current waveform or a good voltage waveform, brought about by virtue of the series inductance of R_{s2} (figure 3).

As is the case with the voltage waveform, an ideal $8/20 \mu\text{s}$ current impulse (together with an acceptable voltage waveform) will only be possible if R_{s2} (figure 3) can be fabricated with a negligible series inductance.

5.3 GENERAL PERFORMANCE

At the time of writing, the combination generator had been used for tests which involved sustained impulses up to 15 kA/kV , and no signs of overheating or flashovers have been observed on any of the components.

The duty cycle for a ten kA/kV impulse is approximately 30 seconds, depending on the charging voltage. The generator has produced the odd 21 kA current impulse and 20 kV voltage impulse with no excessive heating or associated problems.

6. CONCLUSION

It has been demonstrated that a 15 kV/kA combination generator can be readily constructed. A good correlation between predicted and measured results has been observed, justifying the design process. Furthermore, a novel method of fabricating low-inductance, high power, wirewound resistors has been presented. The simple method of flame-spraying

tungsten-carbide onto a copper substrate for use as a trigger gap has also been proven. The design and construction of a suitable measuring circuit and an adjustable, automatically triggered spark gap switch has been presented.

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MEASUREMENT OF LAMINATED STEEL PROPERTIES AT HIGH FLUX DENSITIES AND NON-SINUSOIDAL OPERATION

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ABSTRACT :

A measuring system for determining the magnetic properties of laminated steel is described. The system uses a modified Epstein frame, a 10 kW mosfet power amplifier and a computer-based drive system. The system can operate at flux densities up to 2,2 T, alternatively at fundamental frequencies up to 400 Hz, allowing also for harmonic fluxes. Details are given on the modifications to the Epstein frame, the design aspects of the power amplifier and the software program and digital hardware for waveform synthesizing and data acquisition.

1. INTRODUCTION

The 25-cm Epstein square with double overlapping corners, as proposed by Burgwin in 1936, has been the most widely used and only standardized procedure of arbitration for many years [1], characterized by a high degree of repeatability and reproducibility. Standardized test procedures [2] have been set for the purpose of grading magnetic materials on the basis of loss measurements using the Epstein frame method. The current standard specifies test frequencies of 50 or 60 Hz and allows for magnetic flux densities up to 1.5 T on non-orientated electrical steels and 1.8 T on grain-orientated electrical steels.

The availability of modern solid-state switching devices has made it possible to use AC machines in applications where previously only DC machines could be used. It is well known that saturated machines develop odd current and voltage harmonics in the stator and rotor of the machine due to the double-valued and nonlinear BH-characteristics of iron cores. The use of solid-state motor drives has the disadvantage of generating additional voltage and current harmonics in the machine, and thus contribute to the harmonic loading of, and the losses in, the machine. Calculations need to be based on a square law relationship between iron losses and

flux densities at these high values of flux density as measured data is not available. It has however been found numerous times that the actual iron loss in the machine at these high values of flux density is much higher than calculated. It has thus become necessary to obtain the properties of the magnetic material used in the machine at the flux densities and frequencies that is generated in the machine.

A measuring system was developed to obtain the magnetic properties of sheet steel at flux densities up to 2,2 T at 50 or 60 Hz, alternatively at fundamental frequencies up to 400 Hz but then at a lower flux density. The system also allow for flux harmonics up to 4 kHz. A block diagram of the complete measuring system is shown in Figure 1.

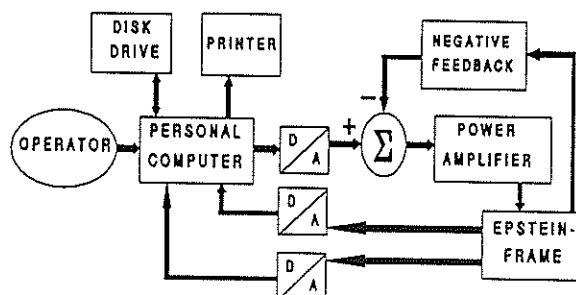


Figure 1. Block diagram of measuring system

A personal computer is used to define the desired flux waveform needed in the test sample. This waveform is transferred via a digital-to-analog converter from the computer to the power amplifier, which is capable of producing magnetic field strengths up to 10 kA/m. The induced secondary voltage of the Epstein frame is integrated and used as negative feedback for the amplifier to ensure that the desired flux waveform is maintained. The primary magnetizing current and induced secondary voltage of the Epstein frame is sampled simultaneously by an analog-to-digital

converter, and the data is transferred back to the computer. The sampled data is processed and saved on floppy disk or hard drive, whereafter the desired graphs are generated automatically.

2. EPSTEIN FRAME MODIFICATIONS

The standard 25-cm Epstein frame specifications and test procedures [2] only allow testing at flux densities up to 1,8 T with sinusoidal flux waveforms. To be able to do tests at the flux densities and harmonic frequencies associated with core and tooth saturation in some electrical machines required that modifications had to be made to the electrical parameters of the Epstein frame design and the drive configuration. The new frame was designed and built by Smit [3]. The physical dimensions of the new frame remains unchanged in order to make use of the standard 305 mm long laminated steel samples.

The frame used consists out of a multi-layer coil comprising 860 turns wound in five layers of 43 turns each, using two parallel 2 mm diameter wires. Taps were made available at the beginning of each layer, which yields turn adjustment possibilities of 172, 344, 516, 688 and 860 turns. A single layer of 840 turns, 0.8 mm diameter wire, was used for the secondary coil. The air-flux compensating inductor was constructed in such a manner that the physical orientation of its secondary coil with respect to its primary coil can be adjusted, thus allowing for the variation of the mutual inductance between the coils without having to change the number of secondary turns.

According to [4] the electrical characteristics measured for the modified frame compared well with that of the standard frame

3. POWER AMPLIFIER

3.1 Design specifications

The power amplifier must be able to supply the necessary current and voltage to the Epstein frame to ensure that the desired flux waveform is generated. By examining the BH-characteristics of several different steel samples it was predicted that a magnetizing force of 100 000 At/m would ensure a flux density of 2,2 T. With the 860 turns on the primary winding of the Epstein frame and the effective magnetic path length of 94 cm it can be calculated that the peak value of current needed is 109,3 A. According to [3] the most severe test condition is at 140 Hz. For a flux density of 2,2 T and a frequency of 140 Hz it can be calculated that a maximum primary voltage of

98,1 V is required.

Figure 2 shows the simulated waveforms of flux, magnetizing current and primary voltage for a saturated coil.

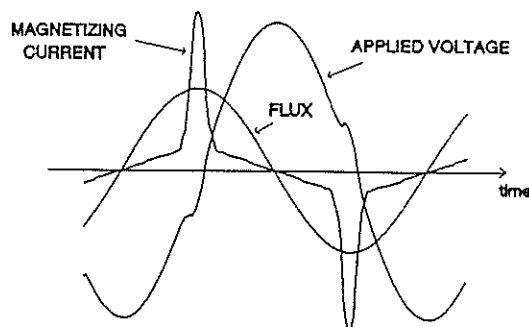


Figure 2. Simulated exciting waveforms

From the figure it is clear that the peak magnetizing current and the primary voltage is approximately 90° out of phase. It is thus possible to reduce the power dissipated in the amplifier output stage by putting series resistive elements in the load path. This not only reduce the voltage across the devices when peak current is flowing, thus reducing the power dissipation, but it also improve the feedback stability.

Doing a test on a sample of steel requires that the flux waveform be generated in the steel, and as soon as transient effects are gone, measurements can be taken. Stability can be reached within 200 cycles, or 4 seconds at 50 Hz. The design specifications for the power amplifier is thus taken as 100 V_{peak} and 110 A_{peak} for at least 4 seconds.

3.2 Resonant inverter vs linear amplifier

A study of several different topologies for the power amplifier revealed that there are basically two different approaches: a linear amplifier or a resonant inverter. The advantages of a linear amplifier are simplicity and easy control, while the main disadvantage is low efficiency. This imply that a lot of power is dissipated in the form of heat, consequently resulting in large heatsinks and a lot of devices. The advantage of a resonant inverter is that, by utilising zero-voltage switching or zero-current switching, the losses in the amplifier can be minimized. This results in a much smaller output stage with less devices. Two disadvantages however are the complexity of the amplifier and its control mechanism, and the fact that the high resonant frequency is superimposed on the output voltage and current waveforms.

A survey was done to determine the feasibility of using a resonant inverter. It was found that the decisive fact would be if a suitable filter could be realized to filter the resonant frequency out of the voltage and current waveforms. The filter specification were set to be a maximum positive or negative gain of 20% in the band-pass region, a high cut-off rate after 4 kHz, at least 3 dB attenuation at 50 kHz, and the impedance of the filter may never be lower than 1 Ω . This imply that the resonant frequency of the inverter will be 50 kHz or higher. After a thorough study of the different filter types [5], a Butterworth filter configuration was chosen due to its oscillation-free gain in both the band-pass and band-stop regions and a sharp cut-off rate.

The transfer characteristics of a filter depends on the load connected to it, which in this case is the Epstein frame. The problem that arise is that due to the saturation that occurs in the coils of the Epstein frame, the impedance of the coils varies drastically during one cycle of magnetization. Figure 3 shows the curve of the inductance of the coils against magnetizing force with a representative sample as magnetic material.

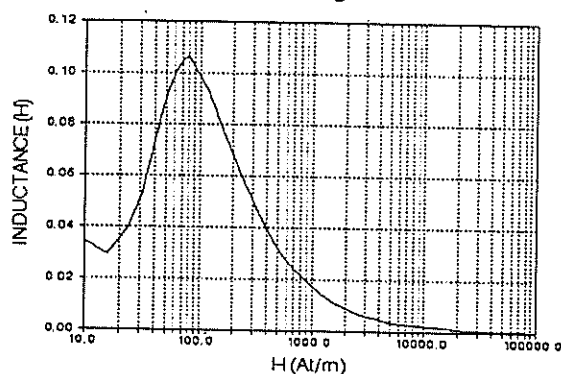


Figure 3. Variation in inductance of coils due to saturation

It can be seen that the inductance of the coils varies by at least an order 1000. This variation in inductance, and thus impedance, has a profound effect on the filter characteristics. Figure 4 (a) shows the frequency response of a normalized 2nd order Butterworth filter, while Figure 4 (b) clearly shows the effect on the filter characteristics of the same filter when the inductance of the load is changed by an order 10.

The best filter configuration was found to be a 2nd order Butterworth, and is shown in Figure 5. The filter element values were determined by simulating the frequency response of the filter for different values of load inductance, ranging from 15 μ H to 120 mH. The effect of each filter element on the frequency response were determined, and then the final element values that satisfied all the specifications were found through

iteration. The final values are $C_1 = C_2 = 15 \mu$ F, $L_1 = L_2 = 7 \mu$ H, $R_1 = R_2 = 0,5 \Omega$.

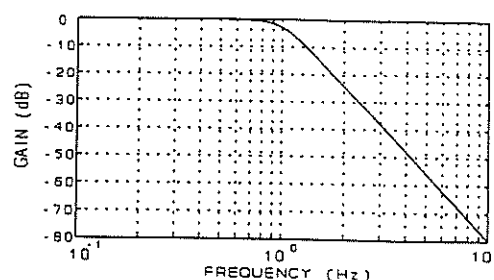


Figure 4 (a). Frequency response of a normalised 2nd order Butterworth filter

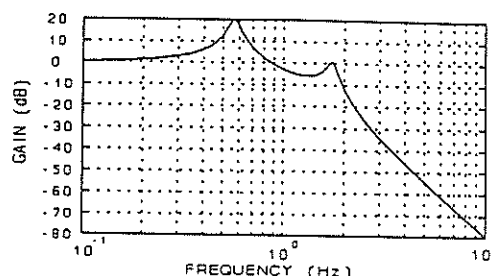


Figure 4 (b). Frequency response of the normalised filter with an order 10 change in load inductance

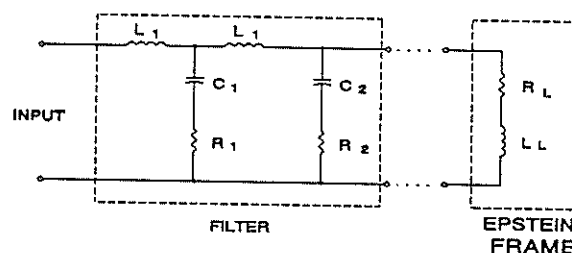


Figure 5. Final filter configuration

In principle it is thus possible to use a resonant inverter as power amplifier, however for the particular application of this amplifier, it was decided that size and efficiency is of no relevance. It is also likely that more difficulties will be encountered with a resonant inverter than a linear amplifier. Due to limited time available it was decided to build a linear amplifier.

3.3 Linear MOSFET power amplifier

The power amplifier consists out the driver stage and the output stage. The driver stage consists out of two operational amplifiers, differential amplifiers and the biasing network. The signal from the D/A card and the feedback signal is amplified by the two operational amplifiers. The difference between

these signals is amplified by two cascaded differential amplifiers. The biasing network ensures that the output devices are correctly biased, and it also incorporates the protection for the output devices.

Due to the relatively high rail voltages and output current rating required, a bipolar transistor output stage as used by Vermeulen [4] is subject to second breakdown problems. It was therefore decided to design the output stage using MOSFETS. It is also well known that MOSFETS can be paralleled without the need for current sharing resistors. A circuit diagram of the output stage is shown in Figure 6. It consists out of four legs, each having seventeen mosfets connected in parallel.

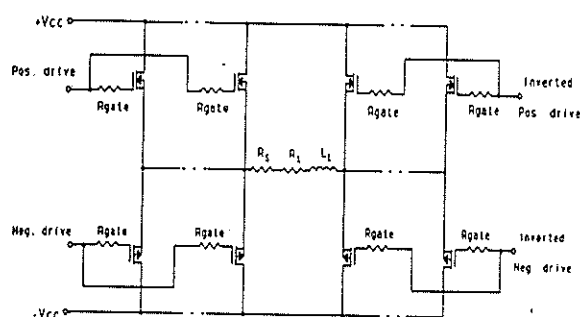


Figure 6. Circuit diagram of amplifier output stage

By using the bridge configuration with complementary pairs of N- and P-channel mosfets it is possible to have the same output power with only half the supply voltage. This lowers the power handling capability of the output stage drastically. Referring to Figure 2 it is clear that the current and voltage waveforms are about 90° out of phase. As previously mentioned the connection of a resistive element in series with the load also reduces the power dissipation in the output stage.

4. CONTROL AND DATA PROCESSING SOFTWARE

A computer program using Turbo Pascal was written to control the measuring system. The program, which is menu driven for easy operation, allows for the operator to define the desired flux waveform. A maximum of 15 harmonics can be specified. The program then generate 500 datapoints describing one cycle of the waveform. The data is downloaded to a transputer-based digital-to-analog converter card with a maximum conversion rate of 250 kHz, from where the waveform is transferred to the amplifier. The

primary magnetizing current and secondary induced voltage waveforms is sampled simultaneously by a 12-bit analog-to-digital converter card. The sampled data is stored under a filename of the operators choice. Graphs of the BH-characteristic, core-loss vs flux density, magnetizing force and flux density is calculated by the program and shown on the screen.

5. CONCLUSIONS

Preliminary results showed that the power amplifier is able to supply the current and voltage required. The menu-driven operation of the system makes it easy to use and the immediate availability of the desired graphs is a big advantage. The next phase of the project will be to evaluate the complete system by correlating measured data to data supplied by a steel supplier.

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ANALYSING A CSI-FED FIELD ORIENTED CONTROLLED INDUCTION MOTOR USING A NEW SIMULATION PACKAGE *CASED*

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Abstract

A model of the current source inverter(CSI)-fed induction motor drive is developed using a new variable speed drive simulation package *CASED*. The mathematical model includes rectifier, inverter and induction motor dynamics. The drive speed control is based upon indirect field oriented control. Simulation results are presented to demonstrate the power of *CASED* as a drives simulation tool and how this control law achieves high dynamic performance from the traditionally low performance CSI-fed variable speed drive.

INTRODUCTION

Numerous simulation studies have been made on the inverter commutation dynamics of the CSI-fed squirrel cage induction motor (SCIM) drive [1],[2]. Stability analysis of the conventional CSI drive control strategies have been conducted where system modelling neglects rectifier and inverter switching dynamics [3]. Akamatsu et al [4] modelled the CSI drive topology under direct Field Oriented Control (FOC) where rotor flux was obtained from flux sensors. Both rectifier and inverter switching operation was accurately modelled. This paper investigates the CSI-fed SCIM drive under *indirect* FOC where the rotor flux is estimated from the stator currents obtained from current sensors. The model developed also accounts for complete drive dynamics of the rectifier input stage, the link filter, the inverter output stage and the SCIM. The simulation model is developed in a new simulation package *CASED* (Computer Analysis and Simulation of Electrical Drives), designed specifically for the simulation of variable speed drives (VSD). The main drive specific features of *CASED* are discussed. Simulation results show not only traditional information such as commutating capacitor voltages, motor terminal voltages, currents and torque but also demonstrates the high dynamic performance achieved from the CSI-fed SCIM under FOC.

CASED : A SIMULATION PACKAGE FOR VARIABLE SPEED DRIVES

To date most investigations of VSD systems through simulation have been such that models specific to one particular drive topology have been developed and equations have to be recast to accommodate different drive topologies. *CASED* overcomes these limitations by providing a general simulation environment which is suitable for analysing the response of a wide variety of complete drive systems [5].

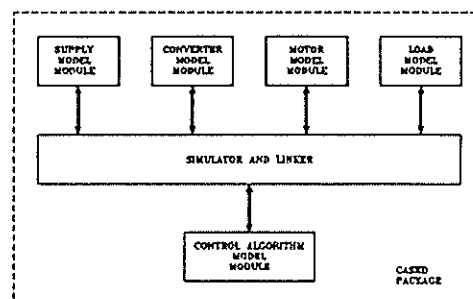


Figure 1 Typical drive system modules interfacing to the *CASED* simulator and linker

CASED has numerous features tailored specifically to the simulation of VSD systems in general. Such features include:

[i]

A piecewise linear state space model of a *general* converter. The converter topology is generated through a graphical interface which then configures the differential equations describing that topology automatically. It is not necessary to predetermine all the possible normal and abnormal conducting patterns of the switching circuits associated with the converter topology.

[ii]

Modular representation of the VSD system. The converter topology is seen as one *module*. The motor model may be generated by the user as another separate *module*. Similarly for the mains supply, the mechanical load and controller algorithm. In this way *CASED* allows the models for each *module* to be developed independently and then linked up to form the complete drive system

(see Fig. 1). The modular concept applies equally to the converter topology where an input stage e.g. rectifier and output stage e.g. inverter can each be generated as separate *module* and then linked up to form the complete converter topology. A library of *modules* may then be developed and subsequently drawn upon to configure any arbitrary VSD system. [iii]

A *powerfull simulator and linker* which combines continuous time simulation with discrete time events typically associated with changing converter switch states and discrete controllers. The simulator accommodates the large difference in times constants often found in VSD systems.

[iv]

Analysis tools where a linearisation option automatically generates the A-matrix of the converter and motor system. Such information is valuable for designing closed loop controllers.

APPLICATION OF FOC TO THE CSI-FED SCIM

FOC : a high performance control strategy

In the DC motor the field (stator) winding is fixed in space so that the field current creates a stationary field flux linkage vector, λ_r . The brushes and commutator ensure that the armature (rotor) current produces an armature flux linkage vector, λ_a , which is in quadrature (decoupled) with the field flux linkage vector at all times. In this way the field and armature flux linkage can be controlled independently to determine the motor torque developed. The field with its large time constant is held constant and the armature current with its small time constant is controlled to allow rapid control of developed torque, T_{dc} .

$$T_{dc} = \lambda_f \times \lambda_a = K i_f i_a \quad (1)$$

In the SCIM, however, the rotor current is induced via electromagnetic induction across the air gap. The rotor flux linkage, λ_r , set up in the rotor due to the induced rotor currents then interacts with the stator flux linkage vector, λ_s , to produce torque, T_{im} . The angle, θ , between the flux linkage vectors, unlike the DC motor, depends on the magnitudes of the flux linkages and the load torque.

$$T_{im} = \lambda_s \times \lambda_r = \lambda_s \sin \theta \lambda_r \quad (2)$$

It is seen from equ.(2) that, if it were possible to control the perpendicular component of stator flux linkage ($\lambda_s \sin \theta$) independently with respect to the rotor flux linkage, the SCIM could be controlled in a similar way to the DC motor. The problem is that in the SCIM not only does the flux or field rotate in the motor at supply frequency but the field and torque producing currents are mixed up in the stator current. FOC "unmixes" the two currents, maintains

constant rotor flux linkage and ensures that its vector is in the correct direction at all times so that the torque producing current is in quadrature with the flux. This is achieved by introducing two axis (D-Q) theory where the three phase stator currents are transformed using the Parks Transform into their equivalent D-Q components and the d-axis aligned with the rotor flux linkage. The result is shown in Fig. 2 with the torque defined by:

$$T_{im} = \lambda_{qs} \lambda_r \quad (3)$$

where:

λ_{qs} = q-axis component of stator flux linkage.

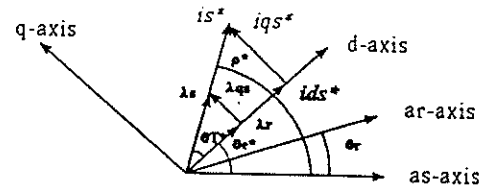


Figure 2 Vector diagram showing the principle of FOC

The d-axis current is used to maintain constant rotor flux and the q-axis current used to control the torque. It can be shown that this orientation is achieved under a very specific condition of slip frequency, $s\omega_e$, given by:

$$s\omega_e = \frac{R_2 i_{qs}}{L_{22} \lambda_{dr}} = \frac{1}{\tau_r} \frac{i_{qs}}{\lambda_{dr}} \quad (4)$$

where:

R_2 = rotor resistance

L_{22} = rotor self inductance

τ_r = rotor time constant

ω_e = stator input frequency

FOC of the CSI-fed SCIM

Fig. 3 shows the complete FOC algorithm in block diagram form. The CSI-fed SCIM has the inherent advantage that being current fed allows the magnitude of the required stator current vector, i_s^* , to be defined indirectly by the magnitude of the DC link current, I_d . The closed loop DC link current regulator of the controlled rectifier input stage holds the DC link current, I_d , to the demanded value, I_d^* . The angular position of i_s^* , ρ^* , can then be established by the inverter output stage. The CSI has six distinct switch states available on the inverter each of which sets up a particular stator current vector in the SCIM (see fig. 4). Such discrete operation presents a problem in the ability to create the desired angular position of i_s^* . One solution to this problem is a technique where an angle compensation regulator is introduced [6] which ensures that the time average of the actual angular position of i_s corresponds to the desired angular position, ρ^* . The error in the angular position, used

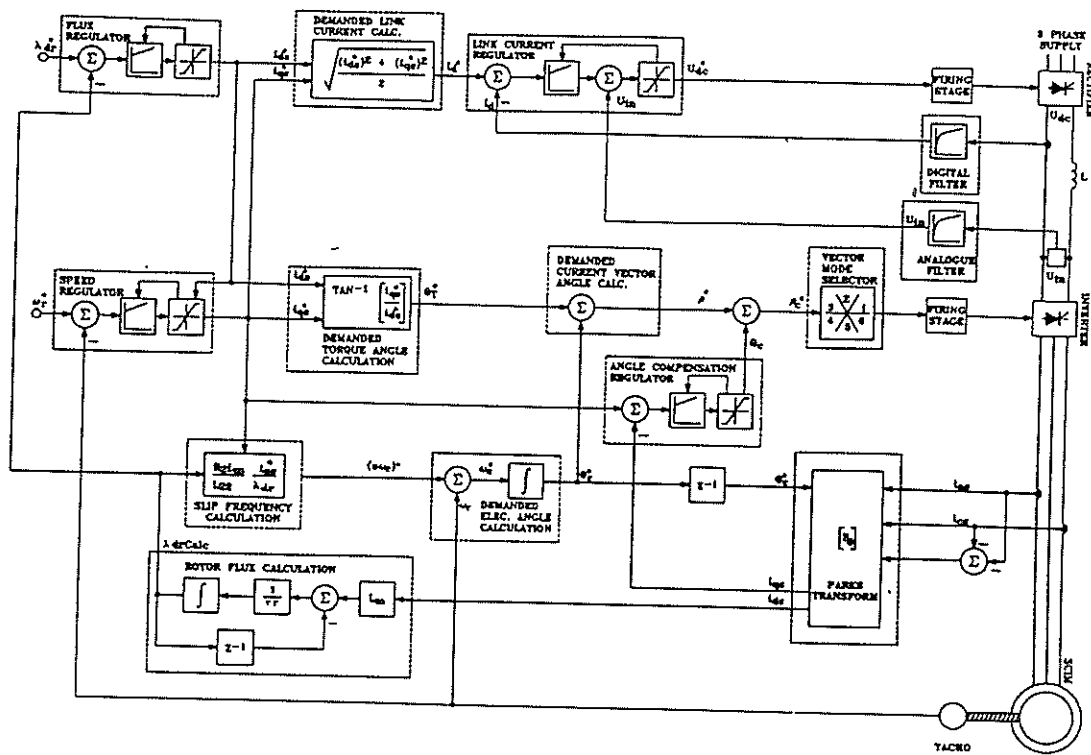


Figure 3 CSI-fed SCIM under FOC : control block diagram

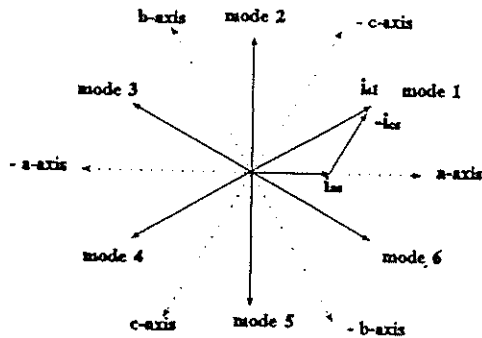


Figure 4 Current vectors for each mode of the CSI

by the angle compensation regulator, is reflected in the error between the actual torque current component, i_{qr} and the demanded value, i_{qr}^* . The output of the regulator is seen as compensating angle, θ_c , which is added to the demanded angular position, ρ^* . At low speeds the operation of the angle compensation loop is seen as PWM operation, where the inverter switches alternately between adjacent modes to create the correct time average angular position. At higher speeds the compensation through time averaging is achieved over numerous modes compensating not only for errors due to the discrete modes of the inverter but also for those errors caused by commutation delay. The magnitude of the rotor flux is evaluated using equ.(5) which

assumes correct FOC conditions i.e. $\lambda_{\psi} = 0$.

$$\frac{\lambda_{dr}}{i_{ds}} = \frac{L_m}{p\tau_r + 1} \quad (5)$$

The slip frequency required for FOC, $(s\omega_r)^*$, is evaluated using equ.(4). The required rotor flux or d-axis angular position, θ_s^* , is then evaluated by integrating the sum of the rotor frequency, ω_r , and slip frequency, $(s\omega_r)^*$ i.e.

$$\theta_s^* = \int_0^t ((s\omega_r)^* + \omega_r) dt \quad (6)$$

The demanded torque angle, θ_T^* , is defined by equ.(7) :

$$\theta_T^* = \arctan \left[\frac{i_{qs}^*}{i_{ds}^*} \right] \quad (7)$$

The required stator current vector position, ρ^* , is then the sum of the torque angle, θ_T^* , and, θ_s^* . The speed regulator produces the demanded torque current, i_{qr}^* , and the flux regulator the demanded flux current, i_{ds}^* .

A Simulation model of the CSI-fed SCIM under FOC

Since *CASED* adopts a technique of modularity the CSI-fed SCIM under FOC requires to be broken up

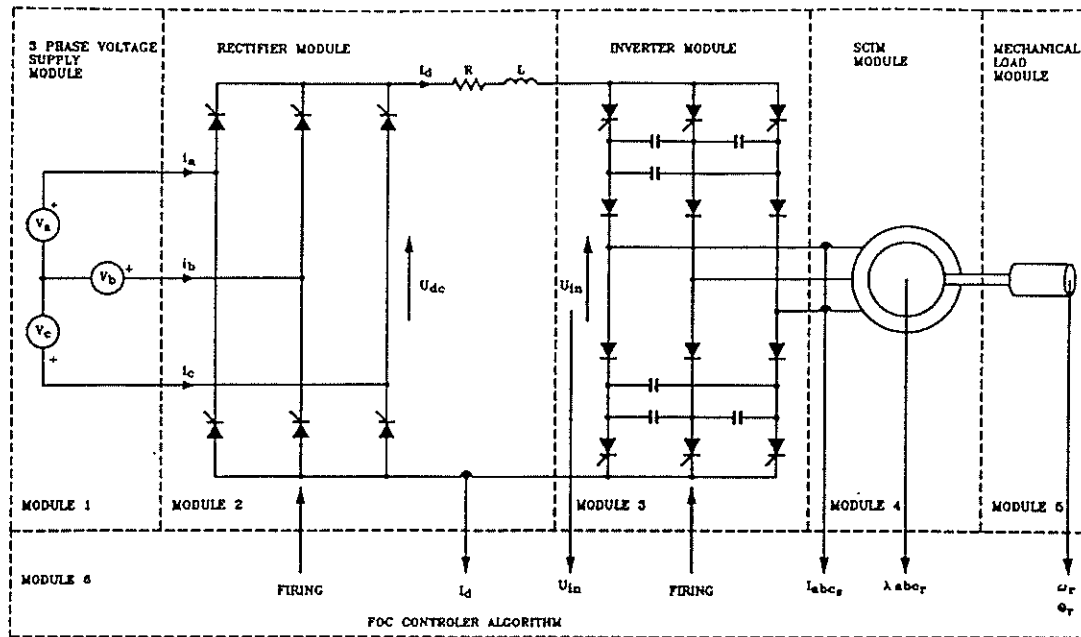


Figure 5 A modular breakdown of the CSI-fed SCIM under FOC

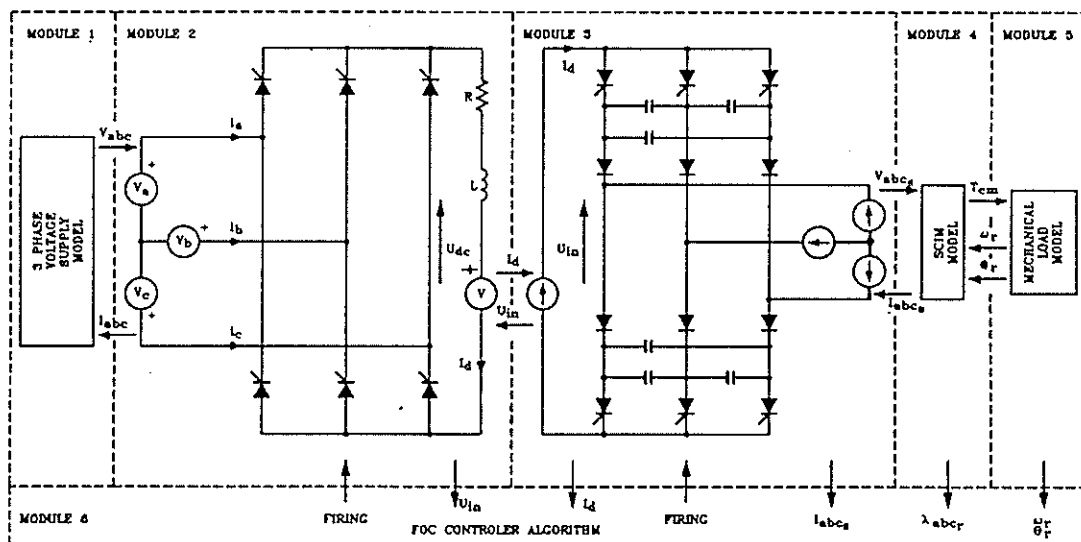


Figure 6 Modular implementation on CASED of the CSI-fed SCIM under FOC

into *modules* suitable for such implementation. The breakdown is arrived at on the basis that each *module* identified may be described by a mathematical model as if it were an entity on its own (see Fig.5). The quantities used as interface values between *modules* are voltage and current between electrical models and torque, speed and angular position between electrical and mechanical models. Using such *CASED* specific conventions, the modular breakdown identified in Fig. 5 may be re-cast into a form suited to *CASED* implementation (see Fig.6). As an e.g. of *module* interfacing *module* 2 is seen as passing the DC link current I_d to *module* 3 which then returns the inverter input voltage U_{in} to

module 2. *CASED* has routines designed specifically to handle such *module* interfacing. The rectifier topology *module* and inverter topology *module* are set up as the networks shown in Fig.6 through a graphical interface, *NETGEN*, which configures the piece-wise linear differential equations of the topologies automatically. The SCIM, mechanical load, and FOC controller algorithm models are coded as *modules* using standard interface routines provided in *CASED*. The supply model is a standard library *module*. These *modules* are then linked together through a another menu driven user interface, *CREATE*, which generates the complete linked simulation model of the CSI-fed SCIM drive

under FOC.

SIMULATED DRIVE PERFORMANCE

The simulated response of the system to a step in speed reference from 0 to 50 and then from 50 to -50 mechanical rad/s is given in graph 2(a)-2(i). Improved dynamic response is clearly seen and the action of the angle compensation loop is evident at at low speeds where PWM operation exists. FOC exists since λ_{dr} is being held at rated value of 1.2 webers while λ_{qr} is being held to zero. Graph 1(a)-1(c) show in finer detail the characteristic waveforms associated with CSI-fed SCIM drive.

CONCLUSION

This paper presented new technique of modelling the CSI-fed SCIM drive, using a new simulation package developed specifically for the analysis of VSD systems in general. The mathematical model accounted for rectifier, link filter, inverter and induction motor dynamics. Indirect FOC was adopted as a technique of speed control. Simulation results demonstrated the power and versatility of *CASED* as a VSD simulation platform and the ability to achieve high dynamic performance from the traditionally low performance CSI-fed SCIM drive.

ACKNOWLEDGEMENTS

The authors would like to thank the Foundation for Research Development and the University of Natal for financial support.

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APPENDIX

SCIM data :

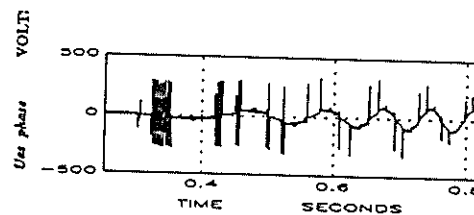
Power	10 hp
Rated speed	1450 rpm
Rated phase voltage	220 V
Rated phase current	14.55 A
Number poles	4

Stator resistance R_1	0.406 Ω
Rotor resistance R_2	0.529 Ω
Stator leakage inductance L_1	4 mH
Rotor leakage inductance L_2	4 mH
Magnetising inductance L_2	125 mH

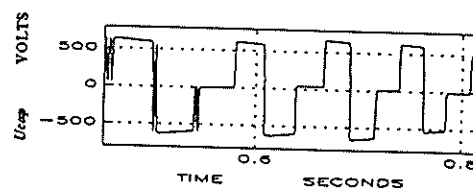
Mechanical inertia	0.5 kgm ²
Frictional coefficient	0.015 Nmsrad ⁻¹

Link inductor data :

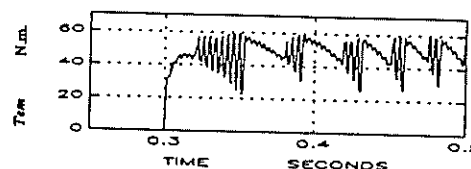
Inductance	150 mH
Resistance	0.4 Ω



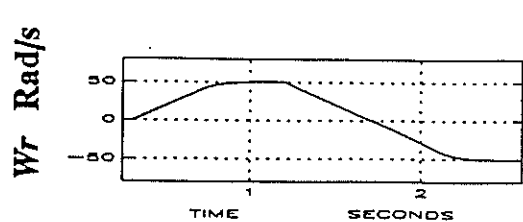
Graph 1(a)



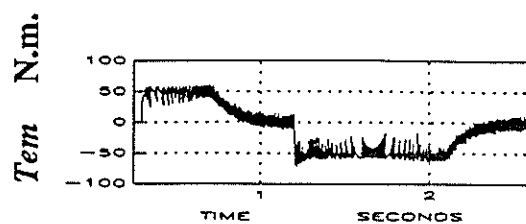
Graph 1(b)



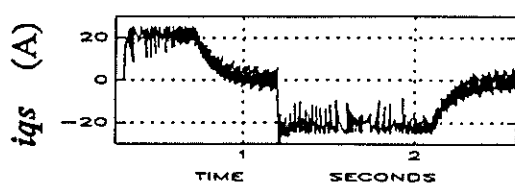
Graph 1(c)



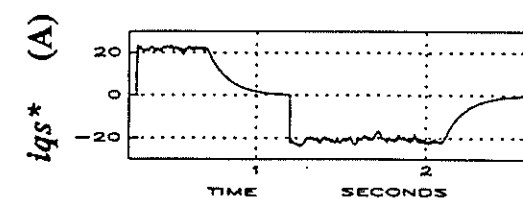
Graph 2(a)



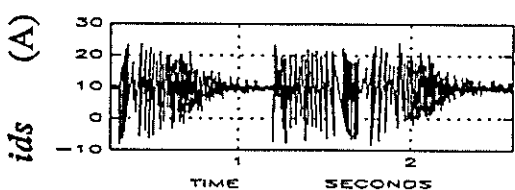
Graph 2(b)



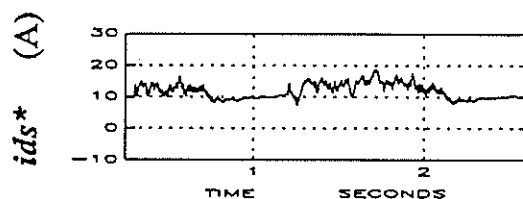
Graph 2(c)



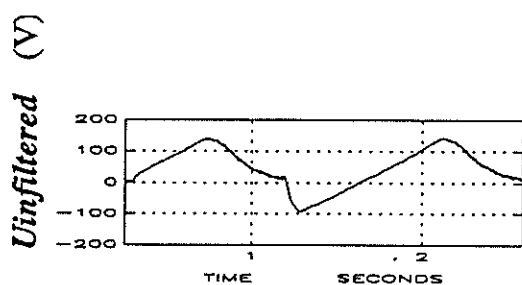
Graph 2(d)



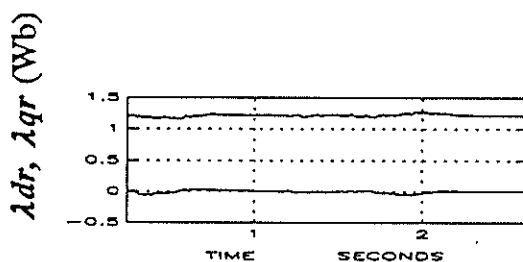
Graph 2(e)



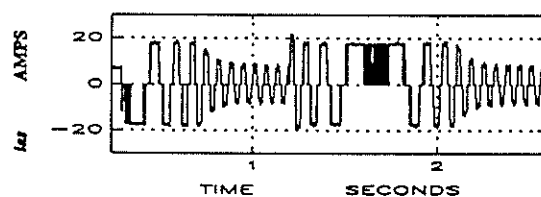
Graph 2(f)



Graph 2(g)



Graph 2(h)



Graph 2(i)

AXIAL VIBRATIONS IN SQUIRREL CAGE INDUCTION MOTORS WITH BROKEN BARS AND INTERBAR CURRENTS

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ABSTRACT

The monitoring of the condition of motors, with respect to broken rotor bars, is typically achieved by either vibration, or current spectrum analysis. Vibration analysis is preferred, in that it is non-invasive, and, unlike current measurement, does not require safety permits for the opening of live substation terminal boxes. However, it is an inconsistent, unstandardised and unreliable procedure. This paper discusses the cause of the axial force in the induction machine due to broken rotor bars, and details a novel method used to predict the occurrence of broken rotor bars.

1 List of Principal Symbols

- $B_{s,s}$ = stator flux density with respect to the stator
 $B_{s,r}$ = stator flux density with respect to the rotor
 e_{bq} = total emf across the rotor bar for harmonic q
 E_{bq} = absolute (rms) emf across the rotor bar for harmonic q
 f = supply frequency (50Hz)
 $I_b(x)$ = bar current distribution along the length of the bar
 I_n = current flowing in a healthy bar
 $\hat{I}_{nq}$ = absolute (peak) harmonic current in a healthy bar
 l = effective (per unit) length of the rotor
 q = harmonic order
 R_c = contact resistance
 s = per unit fundamental slip
 s_q = per unit harmonic slip
 x = axial position along the rotor core from the 'healthy' side
 Z_b = bar impedance
 Z_{bq} = bar impedance for harmonic q
 θ = angular displacement of the stator
 θ_r = angular displacement of the rotor
 ϕ_{bq} = phase angle of the impedance harmonic of the rotor bar

2 Introduction

There are two primary strategies by which the monitoring of the condition of electrical motors is accomplished: the life extension technique, and the life management strategy. The former operates only when a high failure rate indicates that the motor requires monitoring. However, the latter approach manages the entire life of the motor, from arrival at the plant, and requires consistent monitoring of the motor. The life management system is advantageous in that it typically leads to a longer machine life, and enables the systematic capture of relevant data, from which trends can be detailed.

Apart from the information relating to the condition and life of each machine, the life management approach also provides the engineer with information about the scheduling of the each outage. Although condition monitoring warns against possible malfunctions, it must be distinguished from protection, in that it does not protect the equipment against faults.

3 Description of interbar current phenomena

Kerszenbaum and Landy [1] [2] investigated the performance of large, copper bar, squirrel cage induction motors. They demonstrated that when these motors have a broken rotor bar at one end-ring, current nonetheless flows in that broken bar.

In all motors, a three phase current in the stator winding establishes a rotating mmf. It is postulated that, in motors with a broken bar, the voltage induced across the broken rotor bar produces a current which enters the broken bar and distributes itself along the length of the bar. This distribution is given by:

$$I_b(x) = \frac{I_n}{2} \left(1 - \frac{\cosh \lambda x}{\cosh \lambda l} \right) \quad (1)$$

where:

$$\lambda = \sqrt{3 \left| \frac{Z_b}{R_c} \right|} \quad (2)$$

As this current approaches the break, it enters the core and flows tangentially to the adjacent bars.

The current distribution entering the core along the length of the bar is:

$$\delta I_c(x) = \frac{I_n}{2} \lambda \frac{\sinh \lambda x}{\cosh \lambda l} \quad (3)$$

As lambda increases, so the current entering the broken bar approaches the current that would typically flow in a 'healthy' bar. This distribution phenomena can be seen in figure 1.

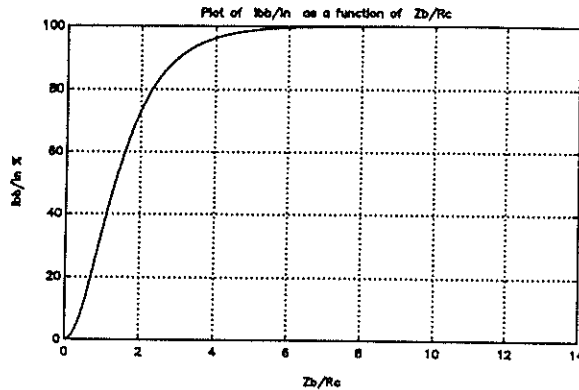


Figure 1: Current Distribution along the broken bar

As the value of lambda increases, so the current entering the core becomes concentrated towards the broken end. This is illustrated in figure 2.

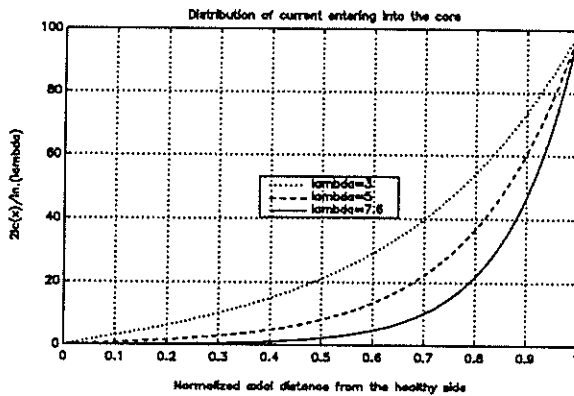


Figure 2: Distribution of current entering the core

From equation 3, it is evident that the core current distribution is dependent on both the bar impedance and the bar contact resistance. First, the bar impedance (Z_b) (and hence lambda) increase due to:

- an increase in the bar length, and/or
- a reduction in the bar area. This reduction in bar area is usually a result of skin effect, when starting a machine with deep or sash bars.

Second, when the bar length increases, the contact resistance (R_c) decreases (which also results in an increase in lambda).

According to Kerszenbaum and Landy [1] [2], large interbar currents flow when the ratio of Z_b/R_c is large. These interbar currents exist when the fault is in the initial stages. Since the machine is continually used, severe burning of the core occurs at the broken or cracked end. This is particularly evident if the machine is repeatedly started. As the burning of the core becomes progressively worse, the contact resistance of the core increases and the ratio Z_b/R_c decreases, limiting the flow of interbar currents.

According to Walliser and Landy [3], these interbar currents reduce the negative sequence currents, hence reducing the imbalance effect. As the flow of interbar currents into the core is reduced, so an imbalance in the rotor occurs and a modulation of current at twice slip frequency is inevitable. The fault can now be detected by analysing the current spectrum at a frequency of $(1-2s)f$ [4]. At this stage, the severity of the fault is such that the whole rotor will need replacing due to burnt laminations and the spread of the fault to two or three bars! This is probably why present motor condition monitoring equipment only detects a broken bar in a large machine having interbar currents when the core is severely burnt and/or several bars are broken.

In the case of smaller machines or die-cast aluminium machines, interbar currents are limited from the outset of the fault. In these cases, analysis of the side-bands in the stator current spectrum is a satisfactory diagnosis technique.

4 Prediction of axial force

The axial force in an induction machine with broken bars is primarily produced as a result of the interaction between the stator produced flux density and the interbar currents. This theory is based on the fundamental equation,

$$F = (B \times i) \cdot l \quad (4)$$

and is seen in figure 4.

Although a force is produced in both the positive and negative axial directions, the net force is only in one direction. This force is oscillatory, due to

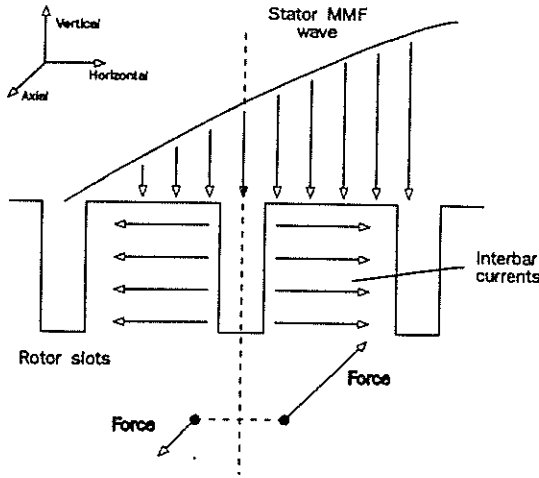


Figure 3: Production of Axial Force

the time varying nature of the current and the mmf waves within the machine.

The flux wave in the air-gap is composed of a 1,-5,7,-11,13 harmonics (Note: the negative sign indicates that the wave travels in a direction opposite to that of rotation of the machine). The flux density of the air-gap, with respect to the stator is given as:

$$B_{s,s} = \sum_{q=1,-5,7,\dots} \hat{B}_q \cdot \sin(\omega t - q\theta) \quad (5)$$

The angular displacement between the rotor and the stator is:

$$\theta = \theta_r + \omega_r \cdot t \quad (6)$$

Referring the stator flux wave to the rotor frame of reference yields,

$$B_{s,r} = \sum_{q=1,-5,7,\dots} \hat{B}_q \cdot \sin(\omega t - q(\theta_r + \omega_r t)) \quad (7)$$

The rotating speed in terms of slip is:

$$\omega_r = (1 - s)\omega \quad (8)$$

Equation 7 becomes,

$$B_{s,r} = \sum_{q=1,-5,7,\dots} \hat{B}_q \cdot \sin(\omega t - q(1-s)\omega t - q\theta_r) \quad (9)$$

The harmonic, per unit slip is,

$$s_q = (1 - q(1 - s)) \quad (10)$$

Equation 9 then becomes,

$$B_{s,r} = \sum_{q=1,-5,7,\dots} \hat{B}_q \cdot \sin(s_q \omega t - q\theta_r) \quad (11)$$

This flux density induces a voltage on each rotor bar, hence by Lenz's law:

$$e_b \propto \frac{dB_{s,r}}{dt} = \sqrt{2}E_b \cdot \cos(s_q \omega t) \quad (12)$$

The current in the bar, I_n

$$I_n = \frac{\sqrt{2}E_b \cdot \cos(s_q \omega t - \phi_{bq})}{Z_{bq}} \quad (13)$$

Rewritten as,

$$I_n = \hat{I}_{nq} \cdot \cos(s_q \omega t - \phi_{bq}) \quad (14)$$

The axial force is given as:

$$F = \left(\sum_{q=1,-5,7,\dots} \hat{B}_q \cdot \sin((1 - q(1 - s))\omega t - q\theta_r) \right) \times \sum_{q=1,-5,7,\dots} \hat{I}_{nq} \cdot \cos(s_q \omega t - \phi_{bq}) \cdot l \quad (15)$$

This Force equation becomes,

$$F = \left(\begin{bmatrix} B_1 & B_5 & B_7 & \dots \end{bmatrix} \cdot \begin{bmatrix} I_1 \\ I_5 \\ I_7 \\ \vdots \end{bmatrix} \right) \cdot l \quad (16)$$

By expanding equation 16 the force components are,

$$F = (B_1 \cdot I_1 + B_1 \cdot I_5 + B_1 \cdot I_7 + \dots \\ B_5 \cdot I_1 + B_5 \cdot I_5 + B_5 \cdot I_7 + \dots \\ B_7 \cdot I_1 + B_7 \cdot I_5 + B_7 \cdot I_7 + \dots) \cdot l \quad (17)$$

By considering the resulting frequency of each term in the equation above, and ignoring the associated phase angle, the following table results:

Frequency	Magnitude
$\sin(2s\omega t)$	$B_1 \cdot I_1 \quad B_5 \cdot I_7 \quad B_7 \cdot I_5$
$\sin(6\omega t - 4s\omega t)$	$B_1 \cdot I_5 \quad B_5 \cdot I_1$
$\sin(6\omega t - 6s\omega t)$	$-B_1 \cdot I_5 \quad B_1 \cdot I_7 \quad B_5 \cdot I_1$ $-B_7 \cdot I_1$
$\sin(6\omega t - 8s\omega t)$	$-B_1 \cdot I_7 \quad -B_7 \cdot I_1$
$\sin(12\omega t - 10s\omega t)$	$B_5 \cdot I_5$
$\sin(12\omega t - 12s\omega t)$	$B_5 \cdot I_7 \quad -B_7 \cdot I_5$
$\sin(12\omega t - 14s\omega t)$	$-B_7 \cdot I_7$

However,

$$B_q \propto \frac{1}{q} \quad (18)$$

$$I_q \propto \frac{1}{q^2} \quad (19)$$

The six times rpm magnitudes thus appear as,

$$F_{6 \times rpm} = - | B_1 \cdot I_5 | + | B_1 \cdot I_7 | - | B_5 \cdot I_1 | \\ - | B_7 \cdot I_1 | \quad (20)$$

The absolute magnitudes of $| B_q |$ and $| I_q |$ are directly related to both the coil span factor, and the distribution factor.

Optimum machine design requires that both factors are as small as possible, to eliminate crawling of the machine. The 7th harmonic complies to this requirement, since the 5th harmonic appears in the braking region. Although $|B_1 I_7|$ reduces the overall vibration level, its effects are small - not only is the winding factor (the multiplication of the coil span and distribution factor) small, but its magnitude is inversely proportional to the harmonic order (7) squared (See equation 19).

5 Experimental verification of 6 × rpm component

A 1.2MW, 4 pole, 3.3kV Pump Drive Motor, exhibited the following vibration spectrum (see figure 4). On inspection, the motor exhibited a crack in the end-ring between two bars.

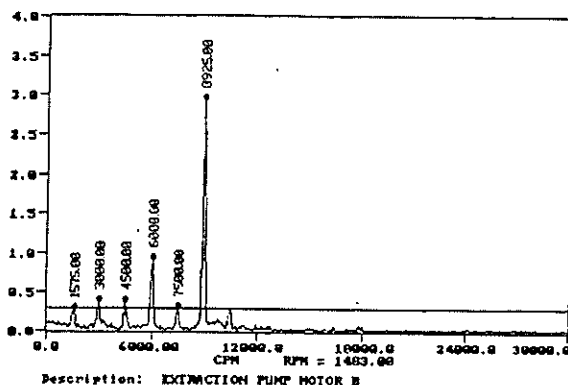


Figure 4: Frequency Spectrum of Axial Vibrations in mm/s

6 Discussion

The axial vibration frequency of the machine depends on the modes of the structure of that machine. It is suggested that an impact test is performed on the machine, in the axial direction, to detect these modes of the resonance. If the modes of the structure lie close to six times the running speed frequencies, then this vibration should be amplified.

7 Conclusion

This paper has demonstrated, both experimentally and theoretically, that a large squirrel cage induction motor, with broken bars and interbar currents,

produces a six times rpm frequency in the axial direction. It is argued that this phenomenon can be utilised as an early warning detection method for broken rotor bars, or for cracked end-rings in a motor with large interbar currents.

Occasionally, large machines with broken bars and interbar currents exhibit a four or a five times running speed component. At present, no explanation as to the cause of these components can be given. Further research into the most suitable measuring points on the machine, as well as the modal resonances of the stator structure, must also be addressed.

8 Acknowledgements

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EXPERIMENTAL ANALYSIS OF THE ELECTRICAL PARAMETERS OF AN OZONE GENERATOR

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1. ABSTRACT

This paper describes the experimental analysis of the electrical operating conditions of two ozone generators in order to find a relationship between important electrical parameters and ozone production. Using digital storing techniques, various waveforms can be stored in a computer for subsequent analysis.

2. INTRODUCTION

A good understanding of the operating conditions is of utmost importance for designing and servicing economical ozone generators. The performance of these machines are evaluated in terms of their ozone production per kVA consumption. For this specific study two commercial ozonators with different designs and performance figures were analyzed experimentally to determine their electrical characteristics and to ascertain electrical conditions for maximum ozone production. Load voltage and current waveforms for both machines were read into the computer through a digital oscilloscope and a GPIB bus. These waveforms were then processed to give parameters like power factor, frequency, amplitude as well as the load power. An equivalent load circuit was also derived from these measurements. These results obtained are very useful, because it is never published due to the commercial value thereof.

3. MEASUREMENT SETUP

When measuring the medium voltage ozonator voltages up to 22kV and currents up to 60A are to be measured. For this purpose a 40kV voltage probe and a 100A current probe were used. These signals were fed into a Tektronix 2430 digital oscilloscope. To link the oscilloscope with the computer, a GPIB interface card and connector cable were used. Minimum requirements of the personal computer are 4Mb RAM and a 286 processor. Tailor made software created in LABWINDOWS was used to control settings on the oscilloscope and to read and store the measured waveforms. An example of a user interface for the software created is shown in Figure

1. After the initial physical connections, all the settings and measurements can be made via the computer. All the waveforms are stored on disc for subsequent calculations. The typical load circuit diagram for the ozonator is shown in Figure 2. The load impedance is represented by a series resistor (R) and capacitor (C). Voltages and currents were measured as shown. An inductor L_s is connected in series with the capacitive load in order to improve the power factor.

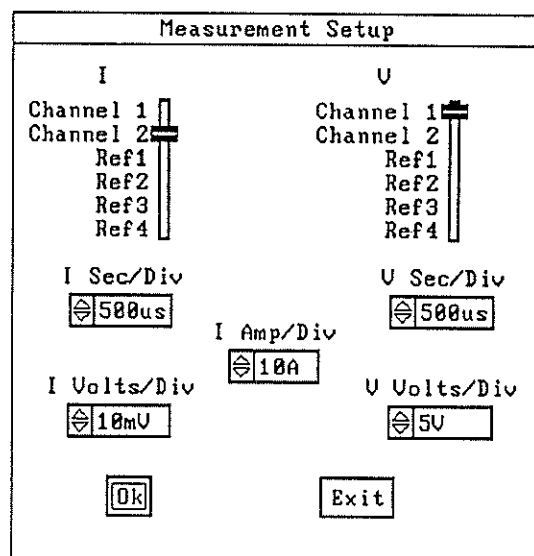


Figure 1. User interface for oscilloscope settings.

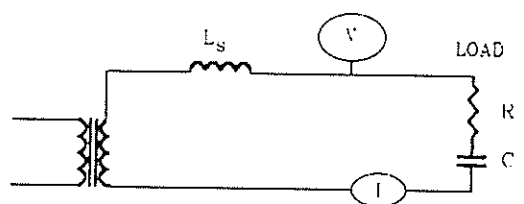


Figure 2. Typical load circuit diagram for an ozonator.

4. EXPERIMENTAL DATA

Typical waveforms acquired at low and high power operation are shown in figure 3a and 3b for machine 1 and for machine 2 in figure 4a and 4b.

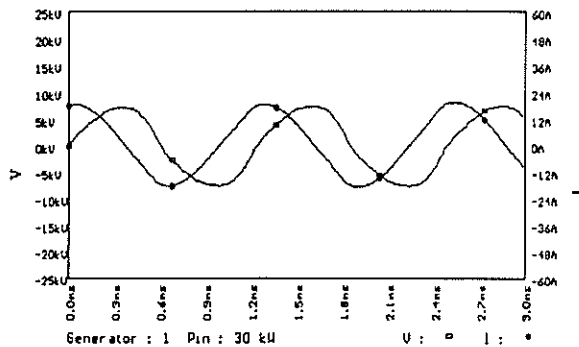


Figure 3 a. Load current and voltage waveforms for machine 1 at low power.

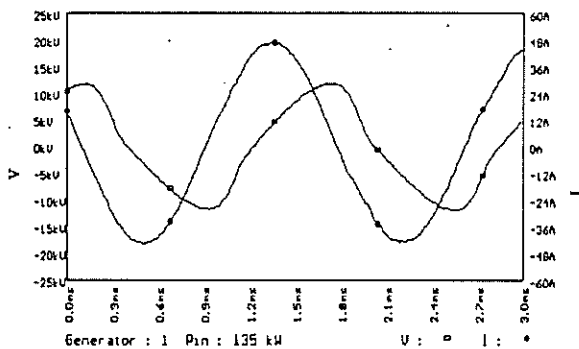


Figure 3 b. Load current and voltage waveforms for machine 1 at high power.

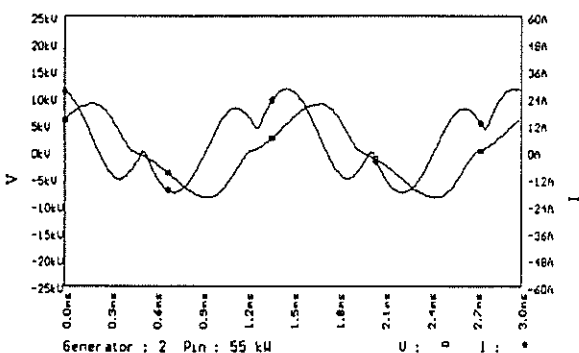


Figure 4 a. Load current and voltage waveforms for machine 2 at low power.

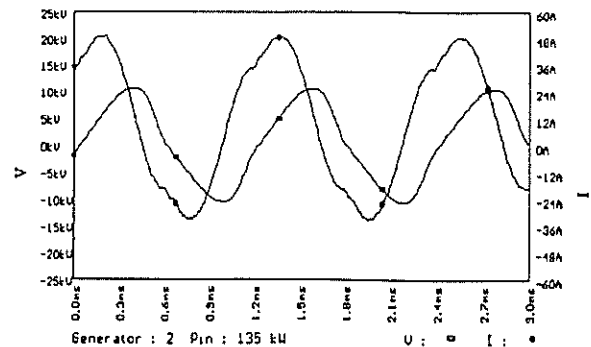


Figure 4 b. Load currents and voltage waveforms for machine 2 at high power.

Frequency increase with power in machine 2, but decrease with power in machine 1. These relationships are shown in figures 8 and 9.

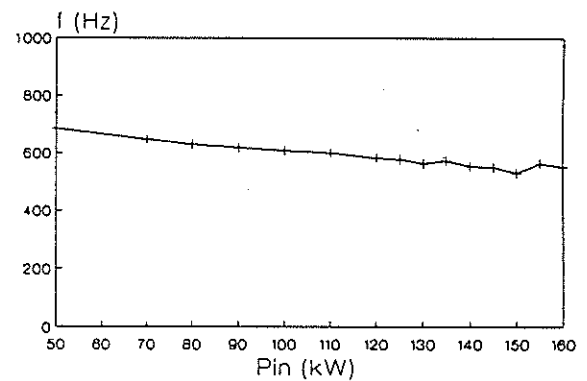


Figure 5 a. Load frequency variation vs input power from machine 1.

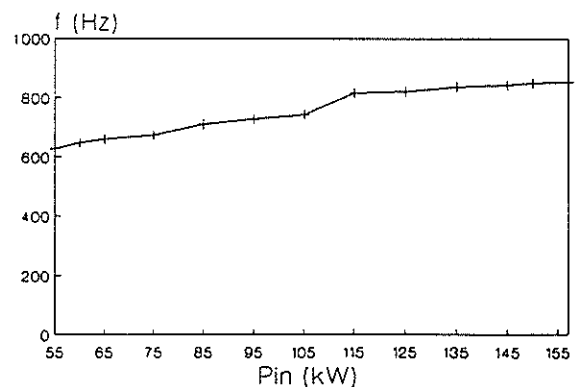


Figure 5 b. Load frequency variation vs input power from machine 2.

5. PROCESSING OF DATA

5.1 Ozone production vs load power

In the introduction it was mentioned that machine 1 produces more ozone per electrical power consumed. The real criterium for determining effectivity of ozone production would be the amount of ozone produced per power delivered to the load (ozonator). These results were calculated from the experimental data using the following equation and available production figures, and are shown in figure 6.

$$P_{AVG} = \frac{1}{t_2 - t_1} \int_{t_1}^{t_2} V(t) I(t) dt$$

Integration for the calculation of power and rms values is performed using the trapezium rule as described by the following equation:

$$I = (b-a) \frac{f(x_0) + 2 \sum_{i=1}^{i=n} f(x_i) + f(x_n)}{2n}$$

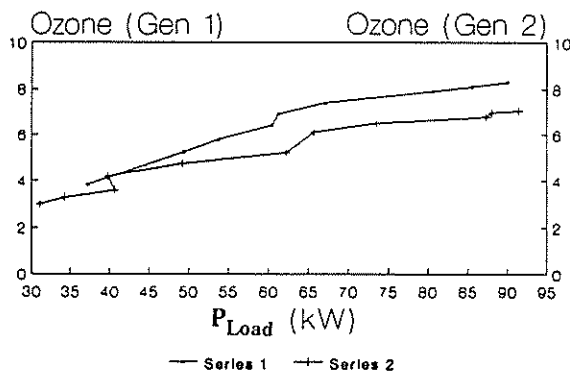


Figure 6. Ozone production vs load power.

5.2. Load impedance vs P_{Load}

The rms values of load current and voltage are calculated using the following two equations, assuming approximately sinusoidal waveforms:

$$I_{RMS} = \frac{1}{t_2 - t_1} \int_{t_1}^{t_2} \sqrt{I(t) I(t)} dt$$

$$V_{RMS} = \frac{1}{t_2 - t_1} \int_{t_1}^{t_2} \sqrt{V(t) V(t)} dt$$

Subsequently the equivalent impedance and series resistance and capacitance values were calculated according to the following relationships:

$$R + jX_c = \frac{V_{rms} \angle \theta}{I_{rms} \angle 0}$$

Values for the resistance were checked using the following formula:

$$P = I^2 R$$

Equivalent values for machine 1 are given in figure 7 and for machine 2 in figure 8.

It is clear that both machines operate at a frequency higher than 580Hz when producing ozone. Machine 2 must be operated from about 55kW, because of the very low power factor and low operating frequency below 55kW.

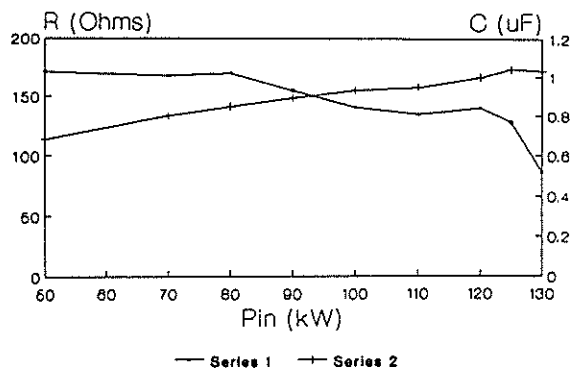


Figure 7. Equivalent R and C values for machine 1.

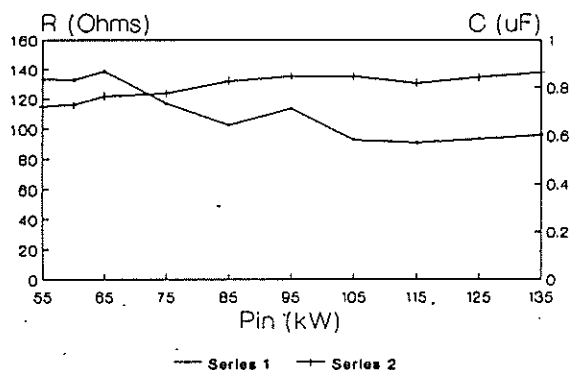


Figure 8. Equivalent R and C values for machine 2.

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6. CONCLUSIONS

As indicated in the measured results, there are some disturbances present in the current waveform of machine 2. As was pointed out by Janowski et al [1], these disturbances can cause a lower ozone production, which is most probably what happens.

This distortion is due to the fact that the high voltage load transformer is fed by an approximate square wave primary voltage. The effect of this condition in the operation of machine 1 is minimized by the fact that the high voltage load circuit is operated close to resonance, resulting in near sinusoidal waveforms. This also confirmed by the fact that the operating frequency decreases while the equivalent load capacitance increases.

Evolutionary Trends in Energy Development in Nigeria

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Abstract

The availability of energy supplies has a profound effect on the quality of human life, organisation of society and economic development. Fossil-fuels and hydro-power constitute the major energy resources used currently in Nigeria. As an oil-exporting developing country, Nigeria will need to depend on her fossil-fuel resources to meet energy demand. However, with the rapid depletion of this resource, the nation is faced with the challenge of providing energy resources to meet future demand. This paper discusses some evolutionary trends in the electricity industry in Nigeria.

1 Introduction

Nigeria is the largest West African country with a land area of about 941849 sq.km, and an estimated population of approximately 100 million (20% in urban areas) an average population growth rate of about 2.5%[1]. Due to large population and efforts to industrialize the country, Nigeria has one of the largest power demand potentials in the African continent.

Electricity is a major resource for energy needs in Nigeria. The present level of development and economic growth achieved in the country, has been brought about by the availability and efficient distribution of electrical

power as well as its safe application.

The National Electric Power Authority (NEPA) was created with the mandate to maintain an efficient, coordinated and economic system of electricity supply for all parts of the country. Nigeria experienced impressive Electricity sales growth in the past. Annual growth rate averaged 18% in the period 1951/52 to 1976/77. The civil war interrupted this growth briefly, but the trend continued resulting in the demand doubling every five years. The oil boom of the 70s played a major role in this development.

As an oil-exporting developing country, Nigeria did not face the devastating impact of the oil price shocks of 1973-74 and 1979-80. This crises had the direct impact of increasing cost of imported oil, indirect impacts of global inflation, high interest rates and the collapse of many commodity prices. Increased earnings on exported oil created a comfortable electric planning environment temporarily.

But this industry under the present negative climate faces numerous challenges such as load-growth assumptions, integrating the emerging and evolutionary renewable energy technologies, political and financial pressures, as well as the enormous cost of infrastructural requirement of rural areas. Nigeria has an installed capacity of 5876MW, an average generating capacity of 3500MW for a present system peak demand of 2362MW.

2 Rural Electrification

The availability of energy supplies has power to transform society and improve on the living standards of people. This is a fact long recognised by governments and policy makers. The government has embarked on the rural electrification project in an attempt to provide electric power to meet the basic needs of the entire people.

When fully implemented, the scheme is expected to enable those in the rural areas (who account for 80% of the population) have access to electricity, and this will accelerate the pace of industrialization. Electric power planning is still based on load-growth projections, while not undermining the constraints of the rural electrification national policy.

The rural electrification project has faced several setbacks in being extended to many remote and sparsely populated areas. These setbacks include: high cost of utility grid extensions and the low initial power demand. These and other factors have brought about evolutionary changes in the general energy mix.

3 Power Planning

As a developing country, Nigeria will need to depend on her fossil fuels and conventional hydro for bulk electric power supplies, especially for industrial loads. Areas of widespread application for the exploitation of renewable energy sources for basic energy supply include autonomous power systems such as decentralized single dwelling and village power supplies for lighting, hot water heating, refrigeration and communications equipment, in addition to direct-drive water pumping systems.

Technologies such as micro-hydro power plants, wind electric generators and photovoltaics will provide a real alternative for grid extension in some remote riverine and sparsely populated areas of the north.

3.1 Renewable Energy Sources

These include hydro, solar and wind. Applications of these sources include those directly related to electricity generation and those not directly related to it. Electricity generation schemes include: micro hydro power plants, wind electric generators and photovoltaics.

Some of the most promising schemes not directly related to electricity generation include applications in agriculture: wind powered irrigation pumps (using direct mechanical drive); rural health care, solar-thermal: for hot water heating, crop drying, food processing and space heating in cooler climates. The Ada Rice project in Anambra and Enugu States, uses solar thermal for crop drying.

3.1.1 Hydropower

Nigeria's largest, most certain and best defined renewable resource for electric systems is hydropower. Efforts are being made to develop fully the technically exploitable potentials in the country since it is a cheap energy source. Beside the Kainji, Jebba, and Shiroro stations, hydropower potentials include Mambilla Plateau, Katsina Ala, Onitsha, Makurdi, Ikom, Zungeru falls[2]. The potential for expanding this generation is high for both large-scale and small-hydro schemes.

Table 1: Hydro Power Stations in Nigeria

Station	Capacity (MW)	Cumulative (MW)
Kainji	760	760
Jebba	540	1300
Shiroro	600	1900

The technology is mature, reliable and low maintenance in character. Although construction costs are high, the zero fuel cost and low maintenance costs make hydro generation of any size generally a competitive investment in the generation mix of an electric

power system.

3.1.2 Photovoltaics

As a supplement to conventional sources of energy, PVs have a range of applications with varying degrees of technical and economic feasibility. The Sokoto Energy Research centre has been established for PV research and applications. Applications include decentralized single dwelling and village power supplies for refrigeration, lighting and communication equipment, in addition to direct-drive water pumping systems.

Industrial applications potentials include sophisticated power supply units for telecommunications equipment, cathodic corrosion protection systems and a number of signalling systems.

3.1.3 Wind Generation

This is accomplished today using modern forms of windmills, with improved aerodynamics yielding conversion efficiencies approaching 50%. Estimation of electric generation potential cannot be put realistically on the same absolute basis with hydro sources. However, at such suitable sites, clusters of wind generators, called "wind farms", may be installed and operated.

At the present state of the technology, taken in competition with conventional generation, wind generation is not generally an economic choice. An area of application in Nigeria for now is in the "stand alone" operational mode, where it provides generation for a load and not connected to the grid. Hybrid pumped storage schemes will find useful application in this regard.

3.2 Biomass

Energy planning including biomass (wood) have been attempted[3], though not with electricity generation. Biomass is closely related

to the traditional energy source used in several parts of Nigeria. However, introduction of recent developments and applications such as combustion, gasification, fermentation, and anaerobic digestion, are finding ready utilization in both rural and urban areas. With fair cost of implementation the units have good prospects for households, farms and livestock.

4 Electricity Growth Rate

During the period 1978/79 to 1992 sales, annual growth rate dropped to approximately 6%. This was due to a combination of factors, some of which include:

Tariff Reform The increase in electricity tariff by October 1979 and 1991, resulted in reduced consumption.

Generation Shortage Non-availability of spare parts to effect maintenance of generating plants also contributed to this decline.

NEPA welcomes the formation of joint ventures with foreign companies to manufacture spares and electrical equipment in Nigeria[4]. There is a big market in Nigeria and the West-Africa sub-region, which is a good opportunity for Electrical Engineers and entrepreneurs.

As an EDC, Nigeria will not pursue a decentralized energy path. Development planners are aware of the difficult challenge faced in some remote areas where utility grid extensions for many rural areas are expensive and may be difficult to build due to rugged terrain or long distances. Diesel and other conventional power options require reliable fuel supplies and service networks, which invariably are expensive to develop and maintain.

Renewable energy is been utilized to meet this challenge in a number of applications, especially in rural conditions, given due considerations to both the technical and economical viability. Moreover, rural electrification

projects suffer from low power factors and hence high fixed charges on investment.

5 Policy Issue

In view of the impending energy crises, measures have been taken in effective management, conservation, exploration, exploitation, and utilization of our depletable energy sources, and gradually reducing our dependence on them while developing more fuel substitutes. Energy policies have been made to determine the types of energy in use, responsibility of research into those needed for the future and the way in which they are to be exploited; taking into account their feasibility, safety, and financial cost as well as social and environmental effects.

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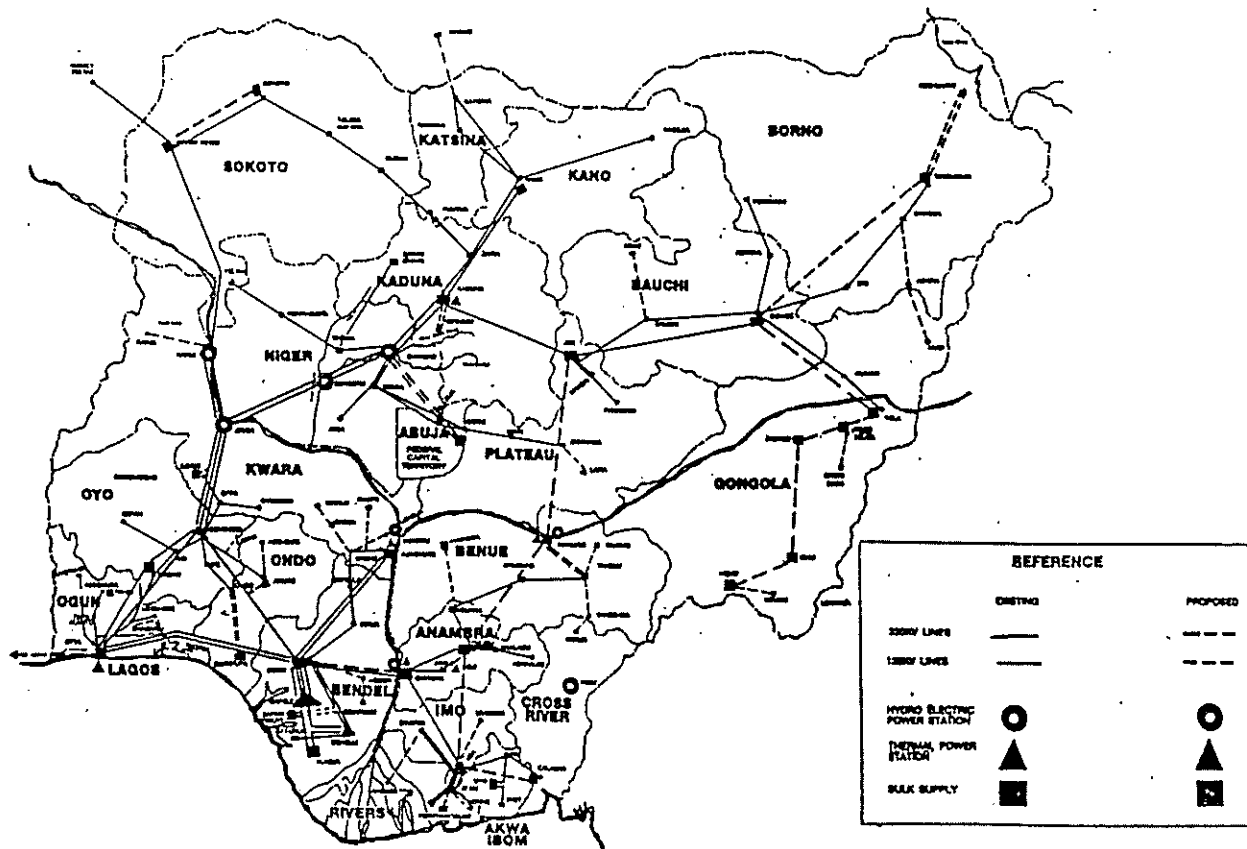


Figure 1: Map of NEPA Grid System

INTRODUCTORY ANALYSIS OF VOLTAGE DROPS IN RESIDENTIAL DISTRIBUTION NETWORKS USING THE BETA DISTRIBUTION MODEL

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In this paper an analytical method for estimating the single phase consumer voltages in a three-phase low voltage distribution system is developed. The starting point is the assumption of a beta model for the statistical distribution of the individual stochastic consumer load currents. The method is further based on a Taylor expansion of the first moment of the consumer voltages with the second moment. Practical implementation of the resultant algorithms is further enhanced by using an approximation for the quantile value of the incomplete beta function and an appropriate curve fit for the Gaussian probability function. The resulting formulation may be used in a computer programme and the results compare well with simulations.

Keywords: voltage regulation, electrical load analysis, statistical modelling, distribution systems.

1 INTRODUCTION

Satisfying the consumer's voltage specification is the most frequently encountered issue during the design of low voltage residential distribution networks. Despite the importance of calculating the expected voltage regulation, the methods generally used by designers are often based on empirical formulations that are not based on rigorous voltage drop analysis. This is because residential loads are stochastic by nature, possessing statistical properties that have been largely misunderstood and ignored in the past. This has been partly due to the lack of good residential load data. In recent years, inexpensive microprocessor-based data loggers have made it possible to redress this problem. Ultimately, the voltage drop question reduces to finding a point estimate of a probable voltage value resulting from a group of consumer loads that may be described by some probability function. Some researchers have attempted to approach the problem with the use of the Monte Carlo simulation method [1]. Such methods involve many repeated simulations, requiring fast computers and may also be affected by peculiarities in the random number generator. They are therefore not directly very practical for design purposes, but are, rather more useful for verifying results from other methods.

Data loggers were specifically developed locally for collecting load current data suitable for use in low voltage design, in particular for voltage drop calculations. The analyses of these data have led to several findings included in publications [2]. The most important development has been the application of the statistical distribution known as the beta function. It has been shown that this two-parameter statistical model is particularly suitable for describing residential

load currents at the time of system maximum demand [3].

The focus of this paper is to derive an analytical method for calculating voltage drops based on the beta distribution model of the load currents. The method has a substantial statistical basis and because the resulting expressions are linear, the algorithms are programmable and therefore useful for practical design applications. Although the analytical approach has been applied to the so called continental distribution system [4] it would be suitable for other systems as well.

2 BASIC NETWORK EQUATIONS

A single section of a four-conductor, three-phase feeder connected to a group of single-phase consumers from the same physical location, is shown in figure 1. This location would typically be a distribution kiosk.

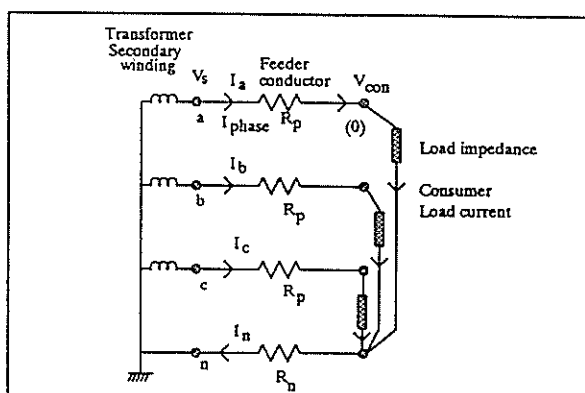


Figure 1 Single section distribution feeder

The individual consumer load currents are assumed to

fit a beta distribution with parameters α and β . Davies [5] analysed the unbalanced voltage drops in feeders using the Gaussian model. Local load surveys have shown that the beta model is more suitable because it is bound by lower and upper limits - as are load currents. Further, it can be either negatively or positively skewed.

Capacitive and inductive reactances are ignored because the cable lengths are relatively short. Referring to figure 1, the neutral current is given by:

$$I_n = I_a + I_b + I_c \quad \dots (1)$$

Taking the a-phase as a reference, the quadrature components of the voltage drop, ΔV , are given by:

$$\Delta V_{re} = I_a(R_p + R_n) - \frac{1}{2}R_n(I_b + I_c) \quad \dots (2)$$

$$\Delta V_{im} = \sqrt{3}/2 \cdot R_n(I_b - I_c) \quad \dots (3)$$

where I_a , I_b and I_c are the magnitudes of the combined consumer currents connected to the a-, b- and c-phases, R_p is the phase-conductor resistance and R_n is the neutral-conductor resistance. Let m_a , m_b and m_c be the number of consumers connected to the respective phases and $Y_1 \dots Y_m$ the individual, scaled load current variates. These variates are independent and identically distributed according to the beta probability density function with parameters α and β and scaling factor C_B . For one consumer per phase:

$$\Delta V_{re} = C_B[Y_1(R_p + R_n) - \frac{1}{2}R_n(Y_2 + Y_3)] \dots (4)$$

$$\Delta V_{im} = \sqrt{3}/2 \cdot C_B R_n(Y_2 - Y_3) \quad \dots (5)$$

and the consumer voltage on the reference phase is:

$$V_{con} = [(V_s - \Delta V_{re})^2 + \Delta V_{im}^2]^{1/2} \quad \dots (6)$$

where V_s is the supply voltage at the sending end.

The aim of the analysis is to obtain an expression for the statistical estimate of the consumer voltage, V_{con} .

3 Statistical analysis based on the beta model

Noting that ΔV_{re} and ΔV_{im} are small compared to V_s , the expected value (or first moment) of the Taylor expansion for expression (6), ignoring terms with greater powers than two is:

$$E(V_{con}) = V_s \left[1 - \frac{E(\Delta V_{re})}{V_s} + \frac{1}{2} \frac{E(\Delta V_{im}^2)}{V_s^2} \right] \quad \dots (7)$$

The second moment of V_{con} , $E(V_{con}^2)$, may be obtained

from (6).

Because the load current variates are assumed to be independent,

$$E(Y_i) = E(Y_j) \quad \dots (8)$$

irrespective of i or j and

$$E(Y_i \cdot Y_j) = E(Y_i) \cdot E(Y_j) = [E(Y_i)]^2 \quad \dots (9)$$

Now, it can be shown [6] that for the beta distribution:

$$E(Y_i) = \frac{\alpha}{\alpha + \beta} \quad \dots (10)$$

$$E(Y_i^2) = \frac{\alpha(\alpha+1)}{(\alpha+\beta)(\alpha+\beta+1)} \quad \dots (11)$$

Thus $E(V_{con})$ can be evaluated in terms of the parameters of the original load current distribution, α and β . Simulations have shown that the resulting distribution of V_{con} can also be suitably modelled by another beta function with parameters of α^* and β^* . Noting that the beta distribution must be bounded by 0 and 1, V_{con} must be scaled. The minimum and maximum voltages are calculated from (4), (5) and (6) by substituting either 0 or 1 for Y_i . This gives:

$$V_{con}^* = \frac{V_{con} - V_{min}}{V_{max} - V_{min}} \quad \dots (12)$$

Thus

$$E(V_{con}^*) = \frac{E(V_{con}) - V_{min}}{V_{max} - V_{min}} = \frac{\alpha^*}{\alpha^* + \beta^*} \quad \dots (13)$$

and

$$\begin{aligned} E(V_{con}^{*2}) &= \frac{E(V_{con}^2) - 2V_{min}E(V_{con}) + V_{min}^2}{(V_{max} - V_{min})^2} \\ &= \frac{\alpha^*(\alpha^*+1)}{(\alpha^* + \beta^*)(\alpha^* + \beta^* + 1)} \quad \dots (14) \end{aligned}$$

4 Development of general algorithms

A general case is considered with the number of consumers on the phases equal to m_a , m_b and m_c and $R_n = k \cdot R_p$. For the general case the maximum and minimum consumer voltages are given by [8]:

$$V_{\min} = V_s - (1+k)R_p C_B ma \quad \dots (15)$$

and

$$V_{\max} = \sqrt{[V_s + .5kR_p C_B (mb+mc)]^2 + .75[kR_p C_B (mb-mc)]^2} \quad \dots (16)$$

The general pattern of analysis followed here is to develop the equations for several sequential cases and then by induction determine coefficients of terms that repeat themselves. The expressions involve combinations of $E(Y_i)$, $E(Y_i Y_j)$ and $E(Y_i^2)$, as given by (9), (10) and (11), respectively. The expressions obtained by induction are given below, without proof [8].

$$E(\Delta V_{re}) = C_{mk1} R_p C_B \alpha / (\alpha + \beta) \quad \dots (17)$$

$$\text{where } C_{mk1} = (1+k)ma - k(mb+mc)/2 \quad \dots (18)$$

$$E(\Delta V_{re}^2) = R_p^2 C_B^2 [C_{mk2} \frac{\alpha(\alpha+1)}{(\alpha+\beta)(\alpha+\beta+1)} + C_{mk3} \frac{\alpha^2}{(\alpha+\beta)^2}] \quad \dots (19)$$

$$\text{where } C_{mk2} = k^2(4ma+mb+mc)/4 + ma(1+2k) \dots (20)$$

$$\text{and } C_{mk3} = \Phi_1 k^2 + \Phi_2 k + \Phi_3 \quad \dots (21)$$

where

$$\Phi_1 = ma(ma-1) - ma(ma(mb+mc) + (mb+mc-1)(mb+mc)/4) \dots (22)$$

$$\Phi_2 = ma(2ma-mb-mc-2) \quad \dots (23)$$

$$\text{and } \Phi_3 = ma(ma-1) \quad \dots (24)$$

In similar fashion

$$E(\Delta V_{im}^2) = R_p^2 C_B^2 [C_{mk4} \frac{\alpha(\alpha+1)}{(\alpha+\beta)(\alpha+\beta+1)} + C_{mk5} \frac{\alpha^2}{(\alpha+\beta)^2}] \quad \dots (25)$$

$$\text{where } C_{mk4} = 0.75(mb+mc)k^2 \quad \dots (26)$$

$$\text{and } C_{mk5} = -3/2k^2 mc \quad \dots (27)$$

Using the expressions developed above a programme may be written to calculate the values of α^* and β^* . The distribution of the consumer voltage, V_{con} , may be derived from expression (12).

In voltage drop calculations for design purposes a point estimate of the voltage distribution is required. This is the quantile value of the consumer voltage at a predetermined level of confidence. Such a value may be obtained from the beta distribution parameters without recourse to a statistical package by using a good approximation for the quantile value [7] given below:

$$x_p \approx \frac{\alpha}{\alpha + \beta e^{2w}} \quad \dots (28)$$

where

$$w = \frac{y_p(h+\lambda)^{\frac{1}{2}}}{h} - \left(\frac{1}{(2\beta-1)} - \frac{1}{(2\alpha-1)} \right) \left(\lambda + \frac{5}{6} - \frac{2}{3h} \right) \quad \dots (29)$$

$$h = 2 \left[\frac{1}{(2\alpha-1)} + \frac{1}{(2\beta-1)} \right]^{-1} \quad \dots (30)$$

$$\lambda = \frac{y_p^2 - 3}{6} \quad \dots (31)$$

Values of y_p may be obtained from standard Gaussian distribution tables. Looking up the values in tables may be eliminated by using an expression that has been found to fit the relationship between probability, p and y_p for values of $p > 0.5$:

$$y_p = -1.294 + 0.7096 \exp(p) - .3718 \ln[\ln(p)] \quad (32)$$

5 Comparison of calculated and simulated results

The results of the analyses for five cases are shown in the table below. Case 1 is for 3 consumers - one on each phase, case 2 for 4 consumers - 2 on a-phase and one each on b- and c- phases, case 3 is for 5 consumers. In cases 1 to 3 $R_n/R_p = k = 1$. In case 4 $k = 2$ and in case 5 two feeder sections of the same conductor size and length, with 3 consumers connected at each distribution point is considered. The consumer voltages were calculated with the aid of a standard spreadsheet programme, using the algorithms derived above. Percentile voltage drops were also determined using Two thousand sample simulations and the percentiles calculated using a computer statistical package, such as Statgraphics®. Since the simulation methods are also subject to uncertainties the agreement between the analytical method described above and the simulated results are acceptable.

Table 1 Comparison of Percentile values

Method	Percentile	Case 1	Case 2	Case 3	Case 4	Case 5
Simulation plus Statgraph.	10%	216.4	213.0	213.7	215.2	211.1
	5%	215.8	212.1	212.8	214.3	209.6
Approxim. Formula	10%	216.6	213.1	213.8	215.5	211.4
	5%	216.0	212.3	212.9	214.6	209.6

6 Conclusion

An analytical method has been developed which calculates the quantile value of the expected consumer voltage in a three-phase, low voltage distribution network. The method is based on the statistical characteristics of the beta model for the load current distribution. The final algorithms are linear and can easily be adaptable to computer software. The method has been compared with simulation methods and found to give good results. It is therefore a useful design tool for low voltage networks.

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AN ELECTRICAL LOAD MODEL FOR EVALUATING THE SYSTEM IMPACT OF INSTANTANEOUS WATER HEATERS

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ABSTRACT

A model for simulating the hot water load of a town is introduced. The hot water load is measured, simulated and then the data is mathematically manipulated to simulate the impact of instantaneous water heaters.

It is concluded that the results of projects such as this one will help the Electricity Supply Industry plan their load growth more carefully and hence apply appropriate marketing strategies today, leading to long-term conservation.

1 INTRODUCTION

According to European standards for water heating systems, the installation of hot water cylinders between the ceiling and the roof and the un-insulated water pipes would not be passed by the building inspectors in Western Europe. From the design point of view, the trend overseas was to create 'wet cells' where bathrooms, kitchens and the like are all grouped together. This reduces the energy and water losses and benefits the home owner by shorter pipe and drainage runs, leading to lower initial costs. Kempster[2] comments on the usage of instantaneous hot water cylinders, where in Western Europe alone 7.5 out of 12 million households are fitted with these heaters.

The aim of this project is to investigate the impact that instantaneous hot water cylinders will have on a South African town.

2 DATA ACQUISITION

The hot water load profile for a whole town has been measured by controlling the domestic water heaters. An aggregated water heating profile was compiled which shows how the system average demand was logged on a minute-by-minute basis for a Tuesday in May 1992. (Figure 1).

The difference between the two distinct profiles in Figure 1 results in a measured hot water load curve for the town. Interestingly it was noted that the hot water load contributed a maximum of 28% of 07h00 to the system load. Although the morning and evening peaks

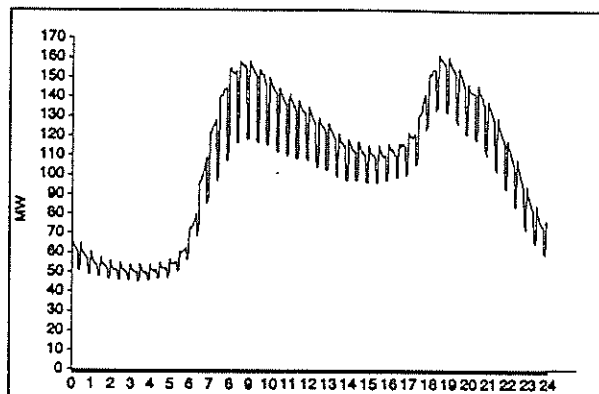


Figure 1 System load when all domestic water heaters are switched off simultaneously each half-hour.

are of similar magnitude at system level, the load due to hot water consumption has a distinct morning peak.

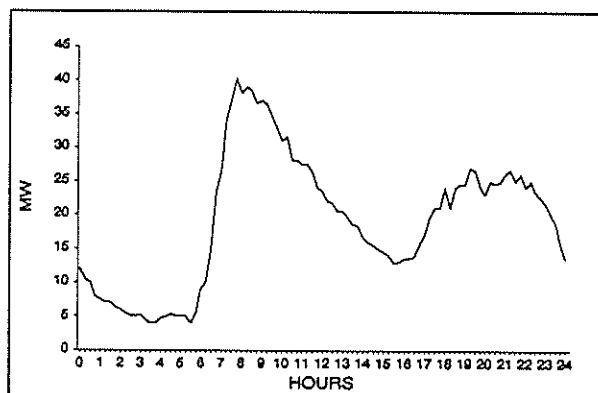


Figure 2 Measured hot water load curve for a whole town for a Tuesday in May 1992.

From the measured data we learned that hot water consumption patterns hardly seemed to vary from day to day, especially during weekdays. It was hard to obtain enough data to follow temperature trends, holiday trends, seasonal trends, etc and repeating the experiment during the winter peak periods proved virtually impossible due to the overloading of substation transformers when the load control is

inoperative.

3 MODELLING THE HOT WATER LOAD CURVE

3.1 PRINCIPLES OF THE HOT WATER MODEL

A probability density function approach developed by N. Beute [1] has been used to model the system level hot water load. This function, also referred to as the hot water model function, is a measure of the probable amount of hot water cylinders that are operative at any given time.

The hot water load at any time t_1 is the load due to the cylinders that have switched on by t_1 , less the load due to all the cylinders that have switched off by time $(t_1 - \delta t_{on})$:

$$P(t_1) = N_{tot} \times Pe \left[\frac{1}{\sqrt{2\pi}} \int_{-\infty}^{\frac{t_1 - t_0}{\sigma}} e^{-\frac{1}{2}x^2} dx - \frac{1}{\sqrt{2\pi}} \int_{-\infty}^{\frac{t_1 - t_0 - \delta t_{on}}{\sigma}} e^{-\frac{1}{2}x^2} dx \right] \quad (1)$$

Pe	=	rating of hot water cylinder
t	=	time in minutes
t_0	=	time at which most people use hot water
σ	=	measure of the width of the distribution
N_{tot}	=	total number of hot water cylinders

This function can be expressed in terms of the Error function:

$$P(t_1) = N_{tot} \times Pe \left[\text{erf} \left(\frac{t_1 - t_0}{\sigma} \right) - \text{erf} \left(\frac{t_1 - t_0 - \delta t_{on}}{\sigma} \right) \right] \quad (2)$$

The energy required to heat M kilograms of water from the inlet temperature to the thermostat temperature can be represented by:

$$Q = M_{H_2O} \times C_{H_2O} \times \delta T \quad \text{Joules} \quad (3)$$

C_{H_2O}	=	Specific heat capacity of water = 4184 J/kg.°C
δT	=	$T_{thermostat} - T_{inlet}$

The time (minutes) taken to heat the water is δt_{on} which represents the time period that the hot water cylinder contributes to the system load and is dependent on the amount of energy required and the rating of the hot water cylinder.

The complete function consists of the summation of a morning curve, an evening curve, a midday curve, and the losses.

$$\delta t_{on} = \frac{Q}{60 \times Pe} \quad \text{minutes} \quad (4)$$

3.2 APPLYING THE MODEL

The probability density function is a symmetrical function. Thus simulating the unsymmetrical morning curve proved more difficult than anticipated and the original model function had to be further refined. A totally different approach was used to identify the load curves. Firstly the town population was divided into two distinct classes, namely: the 'working class' who consume hot water during morning and evening periods only, and the 'domestic class' who consume hot water at random times throughout the day. Secondly to be able to simulate the steep morning increase in the measured hot water load the function had to be refined further by splitting each curve into two sections and defining the ascending slope and descending slope separately.

The losses were based on the SABS Standard 151 and were compared with the measured load curve at approximately 04h00.

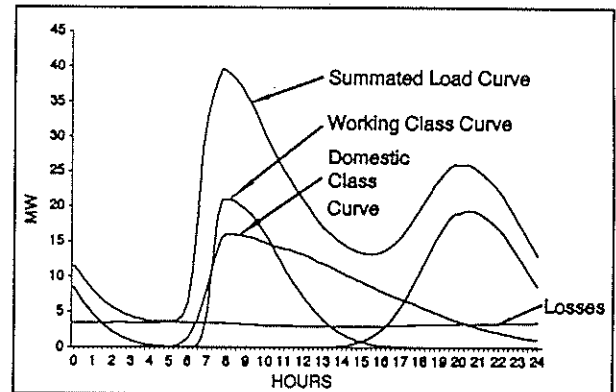


Figure 3 System level hot water load divided into customer class loads and losses.

The three loads as seen in Figure 3 were summated to produce a simulated hot water load curve.

Figure 4 shows the comparison between the measured hot water load curve and that of the simulated curve.

4 MODELLING THE SYSTEM IMPACT OF INSTANTANEOUS WATER HEATERS

An instantaneous hot water heater is an alternative measure to reduce unnecessary wastage. Due to their small dimensions, normally about the size of a shoe box, they can be mounted as close to the draw-off points as possible, reducing pipe losses.

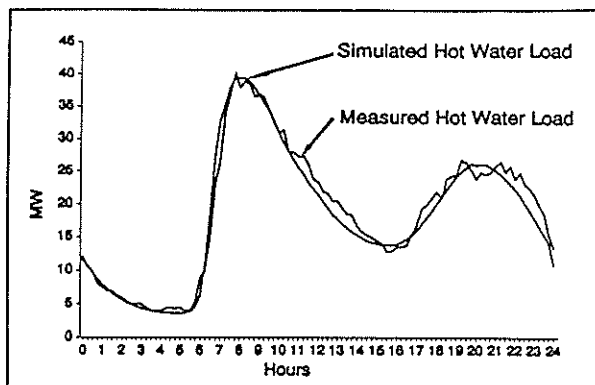


Figure 4 Comparison between the measured hot water load curve and that of the simulated curve.

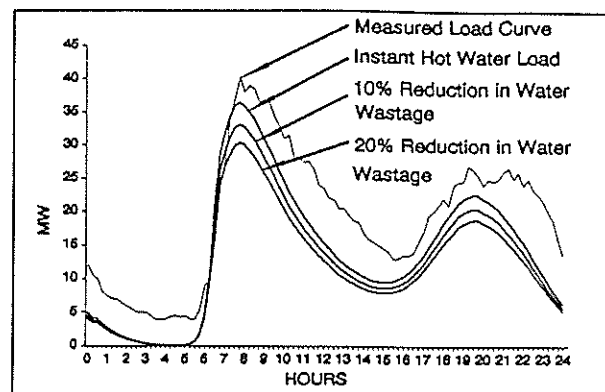


Figure 5 Predicted system hot water load change due to the introduction of instant water heaters and reduced consumption.

A three phase instantaneous hot water heater which is equipped with a 21 kW element was used in the function instead of the conventional 3 kW cylinder. Referring to equation 4, we see that by increasing the element rating, the time the hot water cylinder takes to heat the water consumed is decreased proportionally.

Table 1 Energy losses of the hot water system [1]

DESCRIPTION	WATT	kWh/month
Gross energy/month		300
Standing loss in cylinder	85	61
Standing loss to pipes	30	22
Hot water left in pipes		35
Water passing through pipes		20
Useful energy/month		162

If "wet cells" are formed minimising pipe losses and instant water heaters are used which eliminate the standing losses, the losses can be reduced from 46% to approximately 2%. From equation 3 we see that the amount of water in the pipes is directly proportional to the amount of energy needed to heat this water. Through better planning the hot water load would be reduced further.

Figure 5 predicts the instantaneous hot water load as well as the effects of better planning to reduce hot water wastage.

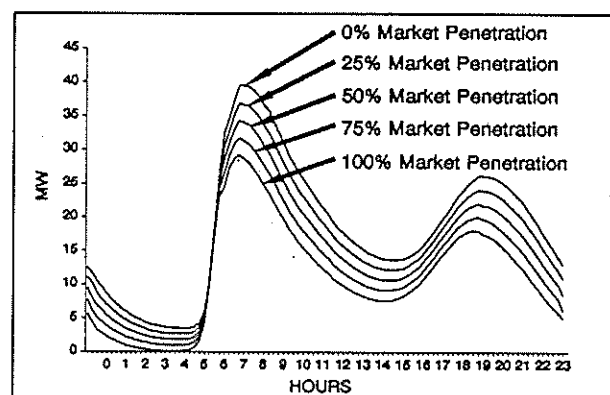


Figure 6 Predicted system hot water load change due to market penetration.

The percentage market penetration of instantaneous water heaters and the influence thereof is shown in Figure 6.

5 CONCLUSION

From the project we learned that implementing instantaneous hot water cylinders not only reduced the peak load but also resulted in a phase shift. The hot water peak as well as the system peak load could decrease by 12 MW if the whole town were to convert to instantaneous hot water heaters.

It is intended that the results of this project, together with several other planned demand-side management projects, will help the Electricity Supply Industry to achieve optimal load growth during the next few years of intensive electrification so that the long-term conservation is not excluded by applying inappropriate electricity marketing strategy today.

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RÉSUMÉ

Lesley Saltmarsh graduated from the University of Pretoria in 1992 with a Bachelors degree in Electrical Engineering. Her interest in demand-side management arose in her final year when she was given a project under the leadership of Prof IE Lane, based on the load management needs of a municipality.

She is currently employed at the City Council of Pretoria, consumer services and energy marketing.

IE Lane is Professor of Energy Systems, Department of Electrical and Electronic Engineering, University of Pretoria. He is presently engaged in research into electricity tariffs and end-use. He is the Director of the Centre for New Electricity Studies at the University.

DEVELOPMENT OF A PERMANENT - MAGNET GENERATOR FOR WIND POWER APPLICATIONS

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ABSTRACT - Low power wind driven generators are normally directly coupled to the wind turbines, while high power generators require their speed to be increased above that of the turbines. It is indicated that directly coupled generators have efficiency advantages, particularly for sites with moderate wind energy potential. An efficient low speed permanent magnet generator, intended for direct coupling to a wind turbine, has been designed, built and bench tested in the laboratory. The objectives of 600 W output at 600 r.p.m., and 90% efficiency at low power output have been achieved. The generator was tested up to an output of 1,6 kW.

1. INTRODUCTION

High speed wind turbines have for a long time been favoured converters for wind energy to electrical energy, due to their high conversion efficiency. Many turbines ranging from the sub-kW size to MW rating have over the last decades been put in commercial or experimental service in many countries. They are particularly attractive options for remote area power supplies. Wind turbines will continue to play a role in providing clean, renewable energy in applications where the economics are favourable.

This paper addresses issues around economic and efficient wind power generator design.

2. BACKGROUND

2.1 Matching of wind data and generator characteristics

When deciding on the ideal turbine-generator set for a particular site, various factors need to be considered. Ultimately the best choice is the machine which will provide energy at the lowest overall cost. Efficient electrical energy generation under all operating conditions is a major requirement, and in this respect the statistical wind data at the operating site needs to be considered.

Figure 1 illustrates the difference in yearly wind speed statistical data for two sites. It can be understood that high conversion efficiency at low and

medium wind speeds is much more important for site A, compared to the requirement at site B, where higher wind speeds prevail for more hours.

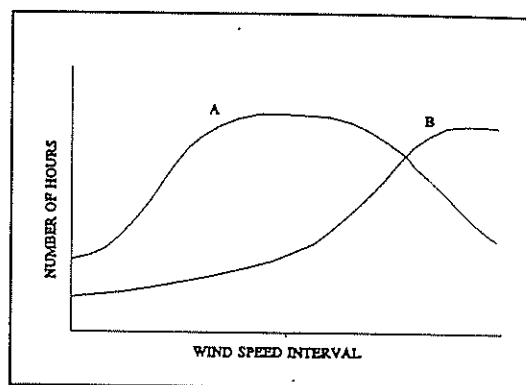


Figure 1. Illustration of yearly wind speed distribution curves for two sites.

This paper addresses the question of an optimally designed generator for the typical site where a large proportion of the annual wind energy is contained in relatively low wind speeds, characterised by graph A in figure 1.

2.2 Speed matching of turbine and generator

High speed wind turbines need to be operated, for their best conversion efficiency, within a limited band of tip-speed ratio, i.e. the ratio of the wing tip speed to the wind speed. Typical tip-speed ratios lie approximately between 4 and 8. It follows that the rotational speed of turbines reduces with increasing power rating and diameter. Figure 2 shows a plot of rated speed vs. power rating for a number of actual turbines of varying types and specifications. (Data obtained from [1]).

The typical operating speed of turbines can be seen to decrease strongly with increasing diameter and power rating.

On the other hand, it is known that electric machines should operate at reasonably high speeds in order to have high power-to-mass ratios and good efficiency. It can be seen that in the low power and speed range

the turbines and generators may be directly connected, but in the high power range a speed increasing coupling mechanism of some sort is required between the turbine and the generator.

Gearboxes, chain and sprocket wheels, V-belt and toothed elastic belt drives are examples of speed-increasing coupling mechanisms which are used.

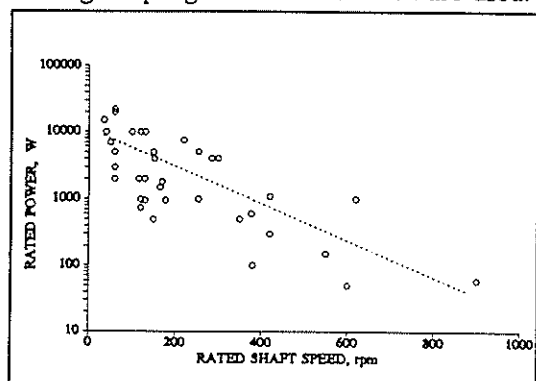


Figure 2. Power vs. speed data of a number of wind energy generators.

2.3 A design philosophy for an efficient medium power generator

There is an intermediate power range where either a direct, or a speed increasing drive may be considered, and the decision will depend amongst others on economic and technical factors, and the statistical distribution of the wind speed.

Two important disadvantages, namely increased initial and maintenance cost, and mechanical complexity, makes speed matching drives undesirable particularly for low power units.

There is a third, inevitable disadvantage, namely drastically increased friction torque on the input shaft.

This is due to the fact that the starting torque of the generator and the friction torque of its bearings, both multiplied by the drive ratio, adds to the starting torque of the turbine. To this is also added the friction torque of the speed increasing mechanism, whether it be a gear box, chain or belt drive.

A low starting torque is the key to effective utilisation of available wind energy in the low wind speed range. This is particularly important at sites characterised by curve A in figure 1.

A single-shaft turbine-generator therefore has a great advantage in such an application.

3. THE DESIGN OF AN EFFICIENT GENERATOR FOR WIND POWER APPLICATIONS.

3.1 Objectives

Based on the above philosophy, the following design objectives were set for a generator to be used in low power wind energy applications:

- direct coupling to the turbine
- permanent magnet field system
- 600 W power output at 600 r.p.m.
- 50 V output voltage at 200 r.p.m. (so as to start charging a 48 V battery at 33% of rated wind speed, i.e. about 3% of rated power)
- three-phase stator winding, full wave rectifier for DC output
- maximum efficiency over 90%, particularly in the low power region

When a generator of this type, which produces voltage proportional to speed, is used for battery charging, it requires a variable ratio DC-DC converter to produce a constant output voltage.

The project only included the design and bench testing of the generator, and not the wind turbine and DC-DC converter.

3.2 Generator design considerations.

It was decided to make use of a standard three phase induction motor frame but the stator winding and rotor were redesigned. Ferrite magnets were decided upon, for reasons of low cost and availability.

A new approach to the analysis of electrical machines resulted in a formula which relates the power-to-mass ratio to speed and the maximum value of operating efficiency [2]. According to this formula, the power-to-mass ratio, when operated at a desired value of efficiency, increases with the square of (flux density x speed x diameter).

It was therefore desirable to obtain a high value of flux density in the airgap. In order to achieve this objective, the length of the stator stack was made half the axial length of the magnets, thereby obtaining in the airgap twice the flux density (0.7 T) of the magnets (0.35 T).

It was necessary to provide pole shoes of laminated magnetic steel, to enable the necessary flux density enhancement between the magnet and the airgap. Pole shoes also protect the ferrite magnets against tooth ripple effects and possible demagnetisation under fault current conditions. A calculation was made of the leakage flux between adjacent poles, and

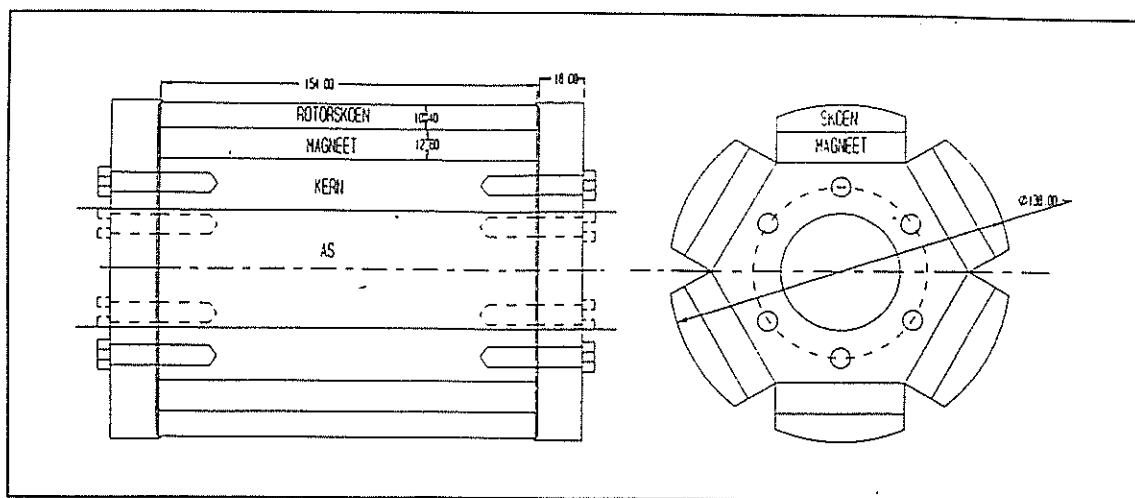


Figure 3. Cross section of the permanent magnet rotor.

this turned out to be less than 3% of the pole flux.

A computer design programme was written which enabled all important physical and performance parameters, like various dimensions, masses of components, magnetic and copper losses, output power and efficiency, to be calculated. The calculation of magnetic losses required available information for magnetic steels at 50 Hz operation, to be analysed to derive general formulae, applicable to varying frequency.

It was possible to vary independent parameters, like the number of pole pairs, airgap diameter, pole arc/pole pitch ratio, etc. and to obtain the physical and performance parameters for each combination of input values.

The best design was that one which gave the highest efficiency at the required power output and speed, within the given constraints.

3.3 Final design results

Some of the important parameters of the final alternator design, when operating at 600 W output and 600 rpm, are as follows:

- no of poles:	6
- conductor losses:	26,5 W
- magnetic losses	3,5 W
- efficiency	95%
- no of slots/pole/phase	2
- no of turns/slot-winding	57
- mass of winding copper	3,3 kg
- slot space factor	48%
- dimensions of magnets (lxwxt)	154x50x12,6 mm
- core stack length	77 mm

The magnets were magnetised in the laboratory, using available facilities.

Figure 3 shows a sectional drawing of the rotor. It consists essentially of a hexagonal, machined mild steel base, with the six magnets and the pole shoes held in place by two brass flanges, bolted to the ends. The rotor was finally covered with fibre glass to obtain a smooth surface.

3.4 Test results.

The generator was connected to a DC motor for bench tests in the laboratory. Current, voltage, input and output power and efficiency were measured over a wide speed and power range. Some results are given.

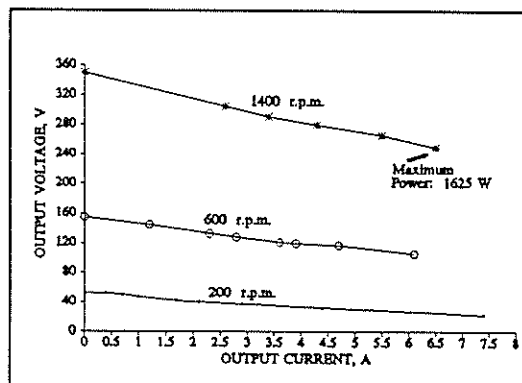


Figure 4. Voltage vs. current curves at three speeds.

Figure 4 shows the output voltage as a function of current at speeds of 200, 600 and 1400 r.p.m. At the maximum current of 6,5 A at 1400 r.p.m., the output power was 1625 W, and the efficiency at this point was 66,5 %.

Figure 5 shows output power and efficiency as a

function of current at 200 and 600 r.p.m. The low speed, low power efficiency can be seen to be 90 %.

The standstill reluctance torque, due to slot reluctance effects, was measured as 0.97 Nm. This torque reduces to an average of 0.19 Nm once rotation starts.

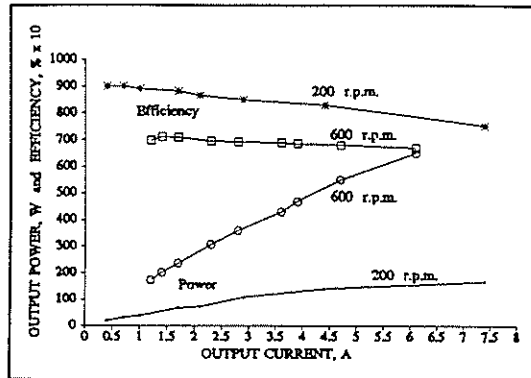


Figure 5. Power and efficiency curves at two speeds

4. PREDICTION OF PERFORMANCE WHEN USED WITH A TURBINE

In the laboratory, the speed and power can be varied independently within wide limits. However, when the generator is connected to a matching turbine in such a way that maximum power is converted at all wind speeds, there will be a unique speed vs. power locus of operation. The variation of voltage with speed (and hence with power) of the PM generator will require in practice that a DC-DC converter be used to provide a constant output voltage. Such a converter may then be controlled so as to provide the maximum available power, using well developed MPPT-techniques.

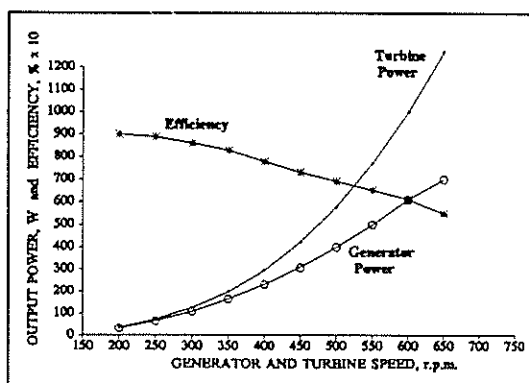


Figure 6. Curves of expected power output and efficiency, as a function of wind speed, for the generator, when controlled for maximum power output.

The expected power and efficiency characteristics of the generator, when connected to a turbine and controlled optimally as described above, have been constructed from the laboratory test data. The resultant curves are shown in Figure 6.

5. CONCLUSIONS

The design, construction and testing of a low speed, permanent magnet generator, suitable for wind energy applications, was successfully completed.

The generator, with target specifications of 600 W output at 600 r.p.m. shaft speed, was designed to be directly coupled to a wind turbine, to generate electrical energy.

The generator has an efficiency of 90 % at low speeds (200 - 300 r.p.m.) and low power (< 100 W). This is a desirable characteristic for the efficient utilisation of wind energy at sites with moderate wind energy potential.

Low values of starting torque (< 1 Nm) and low speed friction torque (< 0,2 Nm) are particularly favourable for its intended application.

The generator has been tested in the laboratory up to an output power of 1,6 kW, achieved at a speed of 1400 r.p.m.

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Mr Jonker was a final year undergraduate student in the Department in 1991, when the work described here was done.

THE CONVERSION OF A SQUIRREL CAGE INDUCTION MOTOR INTO A GENERATOR SUITABLE FOR MICRO WIND TURBINE APPLICATIONS

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Technology Research and Investigations-Eskom, South Africa.

ABSTRACT

Induction generators for grid connected applications and stand-alone applications where they are connected to a resistive load are not new. Based on available literature battery charging application seems to be unusual. The normal route was followed where self excitation was induced by connecting a capacitor bank to the motor. But instead of using a low voltage machine a standard unmodified 380 V, 3 ph motor was used and the voltage stepped down, using a transformer, to a voltage suitable for charging a 48 V battery bank and rectified.

1. INTRODUCTION

In the quest for affordable apparatus suitable for remote area power supply applications the most unlikely options are at times investigated. Sometimes the results are beyond expectations and this was the case where the suitability of the squirrel cage induction motor was researched with the aim to use it as an electrical power source in micro wind turbines used for battery charging.

2. THE PRINCIPLE OF OPERATION

When the squirrel cage induction motor is connected to an AC supply, an alternating current an alternating Emf is induced in the cage rotor very much like in the secondary winding of a transformer. The rotor bars now being a current carrying conductor in a magnetic field starts rotating because of the force of the rotating magnetic field, in the stator, on the rotor bars. Therefore the squirrel cage induction motor can not generate electricity when supplied with rotational mechanical energy alone.

The current which is required to magnetise the induction "generator" must be supplied from the system to which it is connected. If it is connected to an AC network it can function as a generator, as long as it is driven fast enough to produce a positive slip, and receive a magnetising current from the network.

In other words it must be supplied alagging current or an equivalent, namely something which would accept a leading current. These requirements are characteristic of a capacitor.

A capacitor, when connected across the terminals of the induction "generator" forms a parallel LC circuit or oscillating circuit. When the rotor speeds up the residual magnetic field in the rotor cuts the stator winding and initiates small oscillating currents in the LC circuit. At some speed the frequency of the oscillating current reaches a value equal to the resonant frequency of the LC circuit. This speed may vary depending on the amount of residual magnetism left in the rotor and the capacitance tolerance, as shown in figure 1. As this speed is reached the oscillating current between the rotor and capacitor increases rapidly to a value which will allow self-excitation to take place.

Consequently the speed at which the machine will self excite depends on the capacitance as figure 1 clearly illustrates. On the other hand if the capacitance is too low the machine will not start to excite at all or at least not at the speeds at which the prime mover is limited to.

3. VOLTAGE/CAPACITANCE CONTROL

When investigating the relationship of X_C , X_L and the frequency it can be noted that the capacitive reactance would decrease as the speed increases. The result of this is that the oscillating current will increase as the LC circuit made up by the motor and the capacitors get closer to resonance. At which point the impedance (Z) will be equal to the resistance (R) and the current will be at a maximum value after which it will start decreasing if the speed continues to increase. This current is also the excitation current with the result that the terminal voltage would increase in sympathy with the speed.

With a constant speed machine this would not pose a problem but with variable speed applications the need to carefully analyse the operating parameters becomes

critical. When studying figure 1 the tendency would be to use as big a value capacitor as possible to start excitation as early as possible.

The problem with this approach is that at low speeds an acceptable voltage is generated but can increase to above 400 V at 1000 Rpm. As the speed increases the voltage would still continue to rise with 190% of the rated voltage not being impossible. The main problem which can come from this is possible damage to the capacitor banks if their voltage ratings are not high enough as well as the oscillating excitation currents going beyond rated levels.

4. STATIC VAR COMPENSATION

In variable speed applications such as electricity from wind the large variations in voltage for a given capacitance may become a problem. Ideally a machine should have a high value of capacitance at the lower speeds in order to initiate excitation at as low a speed as possible while the capacitance should be reduced at the higher speeds to prevent over voltage.

By increasing and decreasing the capacitance inversely with the speed it is possible to exercise some control over the terminal voltage. This may be achieved by means of an inductively loaded AC/DC converter, a thyristor controlled inductor or switched capacitors. The first two options were not considered because of cost as well as capacitors being the simplest for the purpose.

5. ADVANTAGES

5.1 COST

It is envisaged that a wind-driven generator will be much more cost effective for the following reasons:

- a) The squirrel cage induction motor is a low cost robust machine which is readily available.
- b) An unmodified machine can be efficiency used.
- c) Power is being generated at a high voltage and a low current allowing a wind-driven generator to be placed at the most suitable site regardless of distance. (A transformer is then used at the load end to reduce the voltage to a level suitable for battery charging.)

5.2 VOLTAGE REGULATION / BATTERY LIFE

It is well known that the charge rates have a dramatic impact on a battery's lifespan. As described above by switching more or less capacitance into the circuit depending on the speed the voltage can be regulated. However this would either become very costly if the voltage has to be finely controlled and if not there will still be large fluctuations in the voltage.

When considering that the charge current is dependent on the voltage differential between the generator and the battery and taking note of figure 2 it can be seen that a natural regulation will also take place. As the terminal voltage increases the current will increase accordingly. But it can be seen that the voltage drops as the load increases till the voltage can not be sustained anymore and the voltage and current becomes zero because of the characteristically large regulation of self-excited induction generators. An advantage of this is the added protection it provides the generator under short circuit conditions.

This test was done with a resistive load to illustrate. However when it is connected to a battery it can not reach the point where the generated voltage will be reduced to zero, provided the system design were done correctly. Referring to figure 2 it can be seen that the critical voltage is below the battery voltage. From this it can be deduced that as the generated voltage decreases it will reach a point where the difference in voltage between the batteries and the terminal voltage would be reduced to the point where the current and voltage stabilises at acceptable levels.

To accommodate the fluctuations in wind speeds stepped capacitor banks was experimented with, with the resultant voltage curve in figure 3. If the capacitors are sized correctly and the speed increments selected to match, the regulation of the induction generator can be implemented to limit the effects of these fluctuations to a minimum.

6. CONCLUSION

The above project proved that the squirrel cage induction motor can be a valuable addition to the range of electric power sources for micro wind-driven generators used for charging batteries.

It has a number of advantages but a word of caution however. Most of the design criteria have a mathematical base, but the selection of capacitance values are determined by the performance characteristics of the turbine as well as the electrical characteristics of the generator and an in-depth understanding of the turbine/generator interaction is critical.

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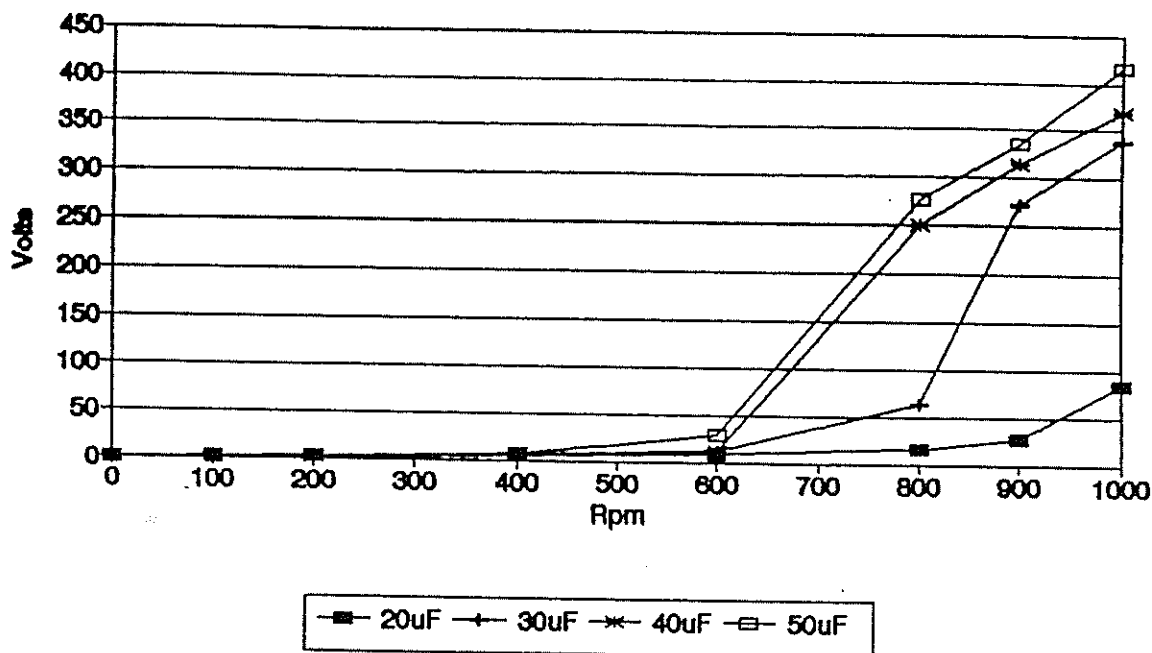


Figure 1.

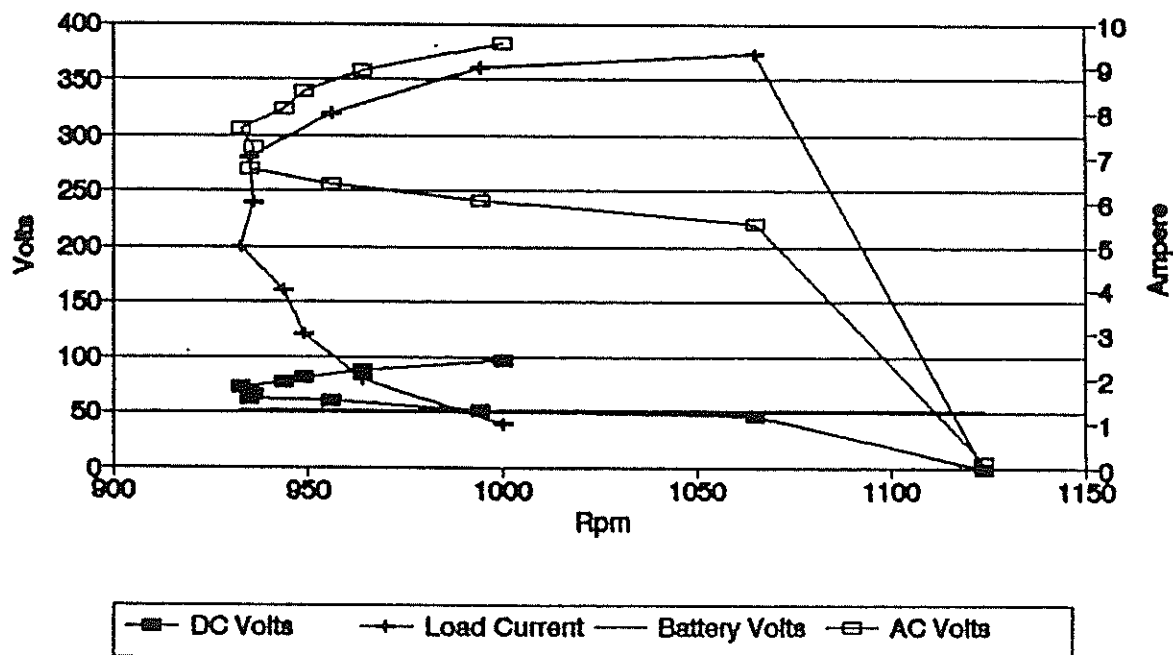
INDUCTION GENERATOR
LOAD CHARACTERISTIC

FIGURE 2.

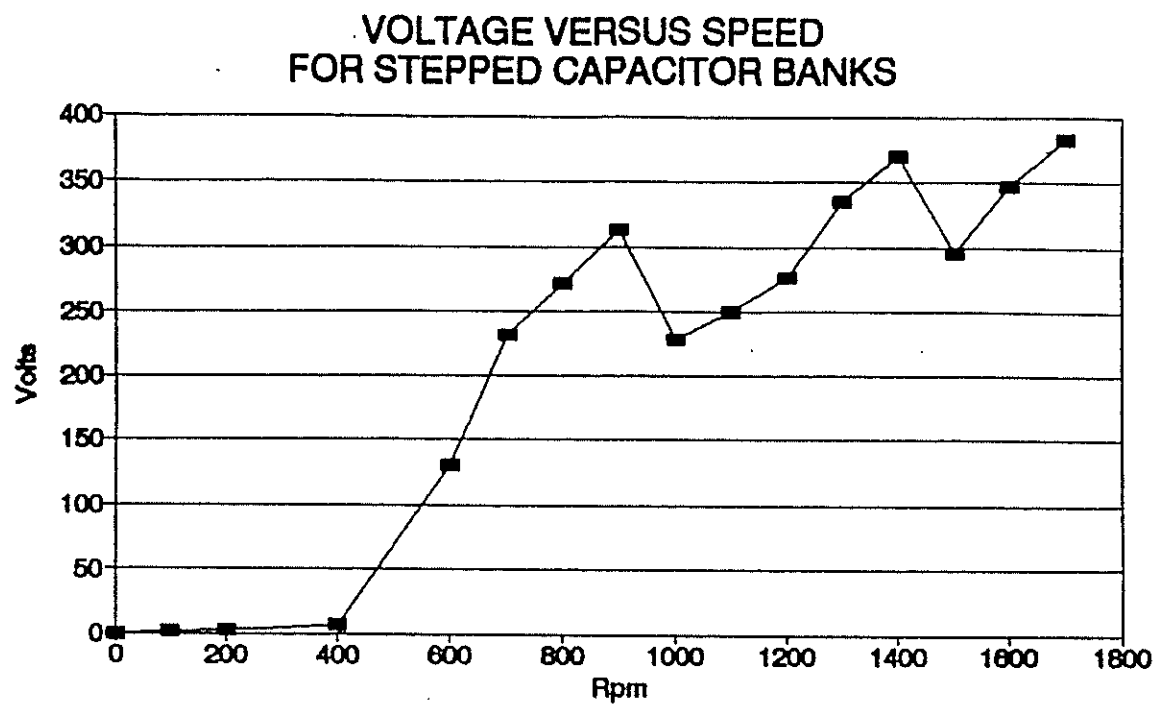


Figure 3.

BEYOND THE VARIABLE SPEED DRIVE

C B McDonnell,* M D McCulloch* and C F Landy**

1. Abstract

Decentralised control is implemented through the integration of the plant parameters and those of the ASD. Consideration is given to the reduction of the total number of measured control parameters within the integrated ASD system to reduce system cost and complexity. This integrated ASD system can now be treated as a black box requiring the minimum amount of information to be passed through the communication link in order for it to be controlled. The digital communication link, made possible by the use of microprocessor controlled ASD's, is investigated. It allows the implementation of a control structure for a multiple drive system as well as drive diagnosis for preventative maintenance.

2. Introduction

This paper introduces a holistic design approach to the integration of adjustable speed drives (ASD) into the industrial process. This involves considering process variables as well as those of the drive when controlling the ASD. Following this philosophy it is hoped that the amount of control information will be minimised while still achieving acceptable control performance. The simulation package CASED is used extensively giving valuable insights to the system. Once this broader understanding of the ASD is achieved, adding intelligence and diagnostic ability becomes a natural extension. The ability to do this is brought about by the extensive use of microprocessors in the control of the drives. At present the digital communication protocol called FIELDBUS [3] is being developed and will prove to be an important base from which to develop more advanced drive communication systems.

By encapsulating the ASD and its associated local sub systems, the overall plant controller can view this sub system as a black box of known characteristics. This abstraction allows for greater ease of use and a more efficient flow of information.

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This paper discusses the aims and scope of the research, the underlying theory and the proposed approach to the problem.

3. ASD Integration Control Issues

As a basis, a vector controlled induction motor drive is to be considered. For such a system speed, flux density and $p(t)$ (the angle of the rotor flux relative to the stator winding reference axis) are typically required to gain full torque speed control of the drive. At present these parameters are used to ensure direct motor control is achieved without regard for the associated process. The plant parameters are also measured and are fed back forming an outer control loop. Alternatively, plant parameters fed directly into the vector controller present a more elegant solution. Clearly this substitution can only take place if the overall quality of control remains within specifications.

As an example, consider a motor controlling a pump feeding into a tank where the level of the tank is to be controlled. The tank level could be substituted in place of the motor speed within the vector control structure. Doing so would still take cognisance of the input and output flow rates, the characteristic performance of the pump and the motor as well as any system disturbances that there may be. Further, this is done without having to measure the motor speed as was previously required.

At present, the process controller is a separate system placed away from the inverter and communication is achieved via a 4-20mA line. A delay is introduced by compulsory filters. If the analogue line is replaced by a more efficient digital link there would still be a delay introduced by the finite baud rate. [1] By incorporating the process control requirements into that of the ASD decentralised control is achieved and this delay is eliminated. Further, since the drive and its associated plant is treated as one entity and not two, less information has to be transmitted, resulting in a more efficient flow.

The research will investigate various plant sub systems to determine the feasibility of incorporating them with the ASD. The investigation will be

performed with the aid of the CASED.[2] This program allows for simulation of the drive, the controller, the plant and their interaction. The overall modelling of the plant and drive is to be done to investigate the form of the new controller. Further, the controllability and observability of the complete system is to be ascertained.

The optimum feedback structure using the minimum number of control variables will be verified using CASED[2]. From this knowledge it should be possible to consider the system as a black box where the minimum inputs are known. The use of the microprocessor to implement the vector control strategy provides a useful front end to this black box. Through the adoption of a digital communication protocol, diagnostic ability and integrated drive control can easily be achieved.

Without the greater flexibility introduced by the microprocessor it would be very difficult to implement the more complex control structures needed to integrate the process and the ASD. It would also be arduous to adapt the individual drive control structure to adhere to the control requirements of the multi drive structure. Further the greater repeatability and reliability ensures that the drive user may remain abstracted from the low level detail of the drive. From a multiple drive system point of view, a reliable and repeatable black box approach is essential if the complete system is to function reliably and accurately.

4. ASD Integration - Communication

A digital communication protocol has been envisaged for the last 8 years, but many difficulties have been encountered in developing an industry standard. This had been due mainly to the conflicting interests of the companies involved in the standards committee, set up by the IEC and the ISA. [3] However a complete FIELD BUS[3] standard has been promised by the beginning of 1994. This will finally provide a firm foundation for the development of multiple drive communication and control systems. Because of the slow progress of FIELD BUS other interim standards have been developed and implemented. These include PROFIBUS[4] developed by SIEMENS AG, FIP a French standard and SERCOS

Whatever the eventual standard is, the advantages of a digital communication link over the existing 4-20mA analogue link will be numerous. There will be an appreciable reduction in the wiring costs. The present analogue system requires a dedicated line to every instrument. Typically a single connection costs \$3000 to a transmitter costing \$800.[3] Digital communications allow for multiple drop nodes on a

single cable. Remote operator control is easily achieved as well as the application of a master hierarchical control strategy. The flexibility of the microprocessor means that little or no motor control hardware change would be required by plant changes and retooling. The altered control structure could simply be reprogrammed at both drive and plant levels. This flexibility can be used to adapt the control structure during the normal operation of the drives. The digital communication structure also allows remote reporting of drive status and diagnostics simplifying the application of preventative maintenance.

FIELD BUS is based on the Open Systems Interconnection model (OSI) and utilises the model's layers 1,2 and 7.[4] Layer 1 is the physical layer, 2 is the data layer and 3 is the application layer. A high transmission rate of 1.5 Mbits/s has already been achieved using the PROFIBUS scheme. This allows up to 63 drives to be addressed in less than 15 ms. [4]

Another communication protocol, called SERCOS, has been designed specifically for servo systems. The protocol uses a data update rate of 4 Mbits/s giving a typical data cycle of 1ms. The nature of the plastic fibre optic links means that the range is limited to 20 m per link which is considerably less than that envisaged by FIELD BUS system.

5. Conclusion

The development of microprocessor control has greatly improved the versatility of modern drives. The discussed plant integration and drive communication are natural developments which will prove essential in an industrial world dominated by the need to be efficient within a flexible manufacturing environment.

The computer simulation package CASED will allow the feasibility of plant and ASD system integration to be accurately determined. The minimising of the required control variables ensures that the integrated system can be treated as an efficient black box requiring the minimum amount of control information.

The resulting decentralised drive system may then be easily integrated using a digital communication structure such as FIELD BUS. The digital structure has many advantages over its analogue counterpart. It allows the implementing of a master control structure as well as allowing drive diagnosis for preventative maintenance.

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Experimental Set-up For Electrical Drives

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ABSTRACTS

The principal aim of studies in a German Fachhochschule is to train the students through close connection to practical experience and a limited time of study. Close connection to practice that means a main part of the time is reserved for laboratory experiments, for work in research and development together with the professor and for practical work in industry. For laboratory experiments modern set-up will be arranged in the laboratories. So happened in the laboratory for electrical machines and drives of the Fachhochschule Mittweida.

In this paper the experimental set-up is described.

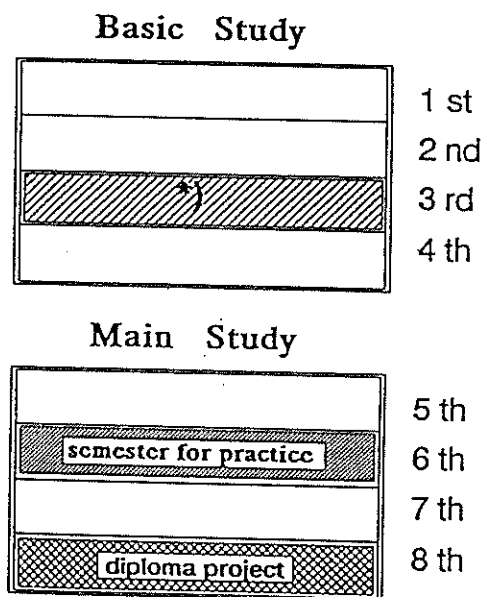
1. INTRODUCTION

The close connection to practical experience and the links to industry at a German Fachhochschule can be seen in the curricula, in research work on different levels and in the field of technology transfer. The students are integrated by means of laboratory experiments, work experience in industry, and by participation in the work of research groups for example during their diploma project.

The availability of modern experimental equipment is an important precondition for successful teaching with practical aims.

2. STRUCTURE OF STUDY

The power engineering students have to pass through four semesters of basic study and four of main study (see Figure 1).



*) not for all fields of study,
semester for practice

Figure 1: Structure of study

Even in the first two years they have to complete laboratory experiments in physics, material science, CAD, electronic circuits, electrical engineering etc. About 15 % of the time is reserved for experimental work.

The main course consists of two semesters of theoretical teaching (5th and 7th semester), a practical training period in industry (6th semester), still one semester for theoretical studies and the diploma project (8th semester).

Especially during the phase of main study the power engineering students have to complete laboratory experiments in nearly all fields of study. In addition to this the best students have the possibility to work in research and developing groups together with the professors. Most of the topics of diploma projects have close connection to aims of practice. The cooperative teaching with other departments ensures that the students get a good overview in bordering disciplines of engineering, business management etc. (see Figure 2).

* Electrical Engineering
* Mechanical Engineering
* Physical Technology
* Business Management

* Microsystems Technology
* Precisions Engineering
* Applied Informatics
* Business Engineer

Figure 2: Fields of study

One field of study of the power engineering students is the topic electrical machines and drives. Because of the resumption of teaching a new laboratory of power electronics, electrical machines and drives was constructed. There are arranged the wellknown basic experiments for electrical machines

- transformers
- DC-motors
- induction motor
- synchronous motor
- special machines (eg. stepper motors and electronic motors are in preparation)

and power electronics with different experiments of converters. The experiments of electrical drives are in preparation. Parallel to this activities a diploma project was started for an experimental set-up with different electrical drives. The aim was to use units which are presently produced in industry and used for

practical purposes.

3. SET-UP AND EQUIPMENT

3.1. OVERVIEW

The set-up (see Figure 3) contains three kinds of drives:

- a DC-converter in connection with a DC-motor
- a frequency converter together with an induction motor and
- a servo amplifier with induction motor.

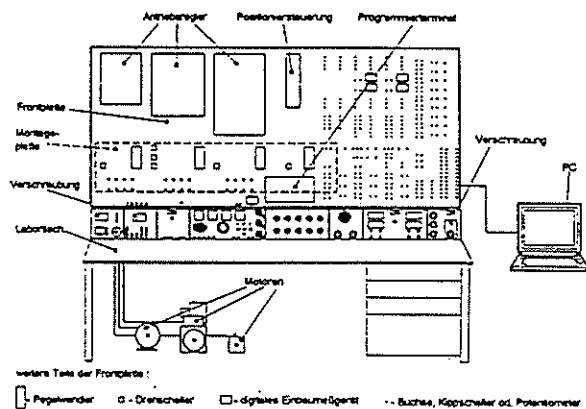


Figure 3: Arrangement of the set-up

All drives can be connected with a free programmable control unit and a personnel computer. All of them are arranged in a vertical board in an analogic way. Finally there are two blocks of arrangement. The first one for the power circuits and their control units, the second for sockets with the in- and output for the control signals.

3.2. THE DC-DRIVE UNIT

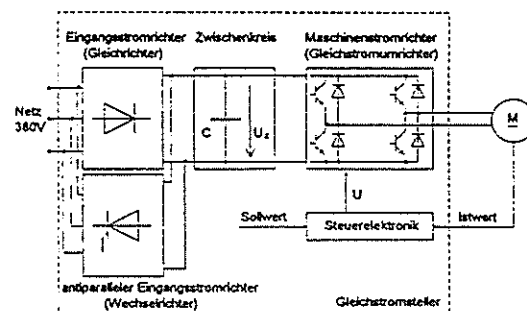


Figure 4: Principle circuit of the DC-drive

The DC-drive input is a DC-rectifier which is parallelly connected to an inverter for retransfer of energy (regenerative braking) and the intercircuit to the DC-chopper (see Figure 4). The DC-chopper allows to control the number of rotations and to change the direction. The speed and the direction of rotation can be influenced by the rated value. This value and further signals are controllable or can be measured at a field of sockets beside the power units. By means of these signals the student can study the function of all units. The principle arrangement of the DC-drive unit is divided in two blocks (see Figure 5), one of the control and power unit and another one for signal sockets.

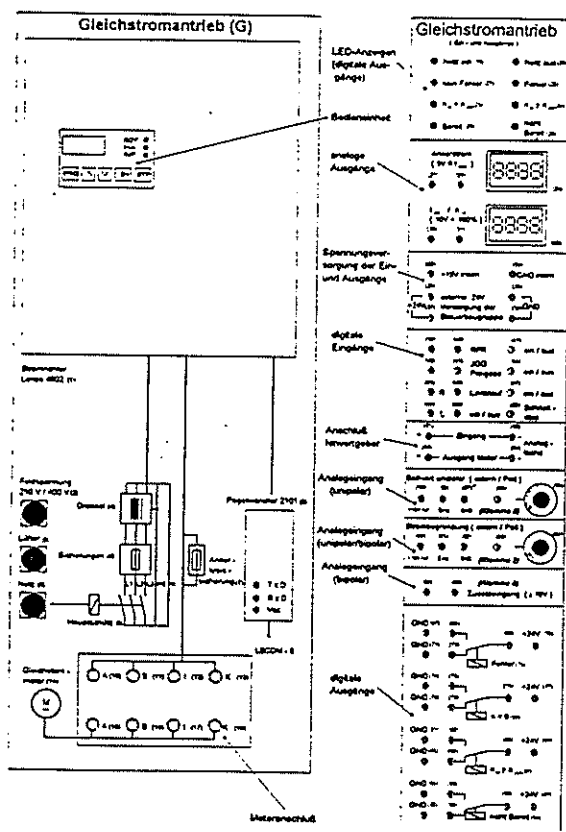


Figure 5: Principle arrangement of the DC-drive unit

For the other drives and the programmable unit it is solved in a similar way

3.3. THE DRIVE WITH FREQUENCY-CONVERTER AND INDUCTION MOTOR

The power unit consists of 4 parts (see Figure 6).

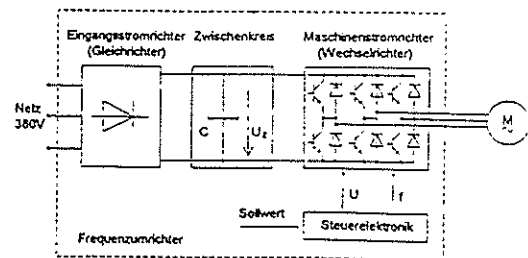


Figure 6: Principle circuit of the drive with induction motor

The input rectifier is connected with the three-phase inverter over the intercircuit. The internal control unit delivers signals of voltage- and frequency-control.

3.4. THE SERVO-AMPLIFIER WITH INDUCTION MOTOR

The servo drive has the same structure as the drive with frequency converter (see 3.3.). In addition to it a chopper is installed, which produces controlled losses over a resistor for braking of the induction motor (see Figure 7).

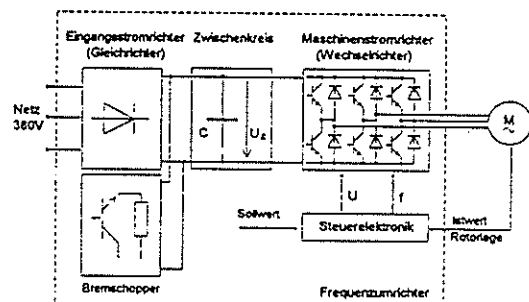


Figure 7: Principle circuit of the servo drive
Further on it has a controller to realize the method of field orientation in rotation control. A position-control unit completes the servo drive.

3.5. PROGRAMMABLE UNIT AND PERSONNEL COMPUTER

It is possible to connect the servo- and the DC-motor to the programmable unit to create values

for way and number of rotation. The programming of the unit takes place on its own terminal or by a serial connection by means of the personnel computer (see Figure 8). Software exists which supports this

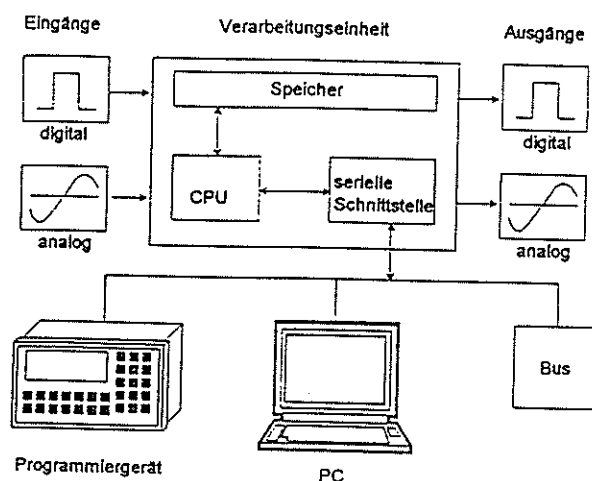


Figure 8: Principle structure of the programmable unit

connection in two directions. The first one is the handling of parameters of the control unit and the second one is the information about the process conditions over the monitor of the personnel computer. The hardware connection is realized by a serial interface (RS 232 and RS 485). There is also a data bank with short informations about the function of all units and sockets arranged at the panel.

4. POSSIBILITIES FOR THE EXPERIMENTAL WORK OF THE STUDENTS

The students get informations about the function of electrical machines, converters, drives and control equipments in lectures and exercises. But the students don't get practical experiences by theoretical teaching.

By using the experimental set-up described above the

students are able:

- to study the structure of drives in practice
- to arrange the circuit connetions by themselves
- to handle the parameters of the control units or converters directly at the service panel of the units
- to use the systems in ON- and OFF-line operation
- to study the function of the bus connection between the different drives and the control units including the personnel computer
- to train the use of a programmable unit in connection with drives and
- to modify the software for changing parameters and handling the process in different ways.

5. Summary

By the construction of the experimental set-up of three different electrical drives the preconditions of practical exercises for the students in the laboratory have been created. The set-up is also delivering the basis for future diploma projects because there are many possibilities for using, extention and improvement. In addition to it the set-up will be used as a reference equipment in the field of technology transfer. The future plan is to complete the set-up above all in the view of mechanical accessories and to improve and to extend the software.

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MV DESIGN FOR RETICULATION SOFTWARE

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1. ABSTRACT

This paper discusses the general considerations and benefits of automatic cable selection, costing and routing for the design of township electrification. The cable routes considered are the medium voltage cables that feed distribution transformers and the low voltage cables that connect consumers. A locally available LV reticulation package that makes use of a CAD system was extended to include medium voltage cable optimisation which results in a significant improvement to this package. It was found that designs could be optimised in terms of cost and performance as the software tool enables alternate designs to be evaluated quickly.

2. INTRODUCTION

In 1991 Eskom announced that the utility was planning to electrify 1.5 million homes over the next five years, in an attempt to make the basic service available to the 21 million South Africans who are still without it[1,2]. The need for advanced design tools, capable of speeding up the design process while maintaining system performance at low cost, for township electrification is thus made apparent by the large demand for electricity.

Different power reticulation packages exist and extensive research in this field is presently being undertaken. On one particular package (WITS University's Computer Aided Reticulation of Townships, CART), development will lead to a powerful design tool that will enable the designer to perform the following:

- locate the positions of consumer terminations, junction points and transformers on plans;
- specify the routing of all medium voltage, low voltage and service cables connecting transformers, junction points and consumer terminations;
- define an allowed set of cable types to be used in a particular design;
- specify ADMD and values for diversity and unbalance factors;
- minimise project cable cost by optimally choosing cable types;

- investigate the effect (cost and performance) on the system of adding cables, deleting cables, relocating design components or changing design parameters[3];
- produce high quality maps, hard copies of calculated results and materials required.

Up to the present, the team working on the above project has concentrated on the low voltage aspect of power network design. This paper discusses a research project that concentrates on creating similar "intelligent" software capable of designing the medium voltage link between the existing medium voltage supply and the transformers feeding the low voltage distribution.

With a system that allows the designer to work with the medium and low voltage sides of reticulation design, the effects of, for example, using less medium voltage cable at the expense of more low voltage cable can be easily investigated. Such investigations and comparisons will lead to "best" designs from a cost and performance point of view as opposed to present designs that have not been optimised.

3. DESIGN CONSIDERATIONS

3.1. CART 1.1

CART 1.1 allows the designer to optimise and try different low voltage design alternatives by providing the option of changing set cable routes, junction point and transformer positions and design parameters. Optimisation routines yield the most cost effective cable cost schemes that provide the consumer with a voltage that falls within statutory limits.

This version of CART is particularly suited to reticulation designs that make use of a more complex low voltage network being fed from a simple medium voltage network. CART 1.1 is presently being used by designers working in this field.

3.2. CART with Medium Voltage.

Addition of medium voltage design tools to enable overall optimisation of MV and LV networks will extend the useability of CART to many more types of reticulation design.

The medium voltage design tools should allow the designer to use any of the following kinds of reticulation methods :

- Three-phase medium voltage cables with three or single phase transformers.
- Single-phase medium voltage cables.
- Three or single phase intermediate voltages (6.6kV or 3.3kV).
- Mixed systems with the appropriate low voltage cabling.

3.2.1. Junction Equations

A thorough understanding and concise statement of relevant medium voltage equations that describe the above scenarios is required in order to make the MV additions to the software. Values for current and voltage can be calculated at each junction (either distribution transformer or medium voltage tap-off point) of the medium voltage network as follows:

3.2.2. Three-Phase Distribution Transformer Reticulation (see Figure 1)

Three-phase medium voltage feeds three-phase distribution transformers either directly off the main medium voltage feeder or from tee-offs from the main feeder. Transformers or tee-offs are referred to as being junctions.

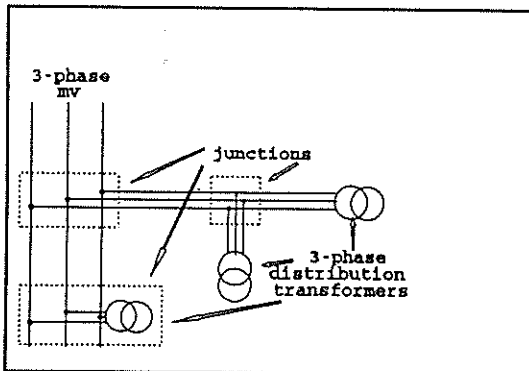


Figure 1: Three Phase Distribution Transformer Reticulation.

The current in the MV cable feeding this type of junction is given by:

$$I_{phase} = \frac{N * I_{max} * CF(N)}{3 * a} \quad (1)$$

where:

- I_{phase} is the mv current per phase in the mv feeder

feeding this junction.

- N is the total number of consumers in the MV cable sub-tree.
- I_{max} is the allowed consumer current.
- CF(N) is the coincidence factor for N consumers.
- a is the transformer ratio.

The per phase volt drop in the MV cable is given by:

$$V_{phase} = \frac{I_{phase} * \alpha_{mv} * L * UCF(N)}{1000} \quad (2)$$

where:

- α_{mv} is resistance of the MV feeder per phase per unit length (given in Ω/km), it is assumed that power factor is 1 as residential loads are generally not highly capacitive or inductive.
- L is cable length in metres.
- N is the number of consumers fed by this feeder.
- UCF(N) is an appropriate unbalance correction factor for N consumers depending on cable and/or transformer configuration.

Should there be a transformer at this junction, the volt-drop within the transformer due to internal impedance must be taken into account. "Most planners ignore the voltage drop over MV/LV transformers which could lead to serious errors especially if the transformer reactance to resistance ratio is low and the percentage impedance is high" [4]. The volt drop can be calculated if the percentage impedance, transformer rating, reactance to resistance ratio, primary or secondary currents and voltages of the transformer are known:

$$Z_b = \frac{V_b^2}{P_b * 1000} \quad (3)$$

$$Z_{sc} = \frac{Z * Z_b}{100} \quad (4)$$

$$X_x = \sqrt{\frac{Z_{sc}^2}{1 + (X_{to}R)^{-2}}} \quad (5)$$

$$R_{sc} = \frac{X_{sc}}{X_{toR}} \quad (6)$$

$$MV_{drop_{trf}} = I_{typ} * a * (R_{sc} * \cos(\theta) + X_{sc} * \sin(\theta)) \quad (7)$$

where:

- θ = power factor angle = 0 as power factor of 1 is assumed for residential loads.
- Z_b , V_b , P_b are per unit base values for impedance, voltage and power.
- Z_{sc} , X_{sc} , R_{sc} are impedance, reactance and resistance values referred to the transformer secondary.
- X_{toR} is the reactance to resistance ratio.
- $MV_{drop_{trf}}$ = volt drop within the transformer referred to the primary side.
- I_{typ} = lv current per phase for number of consumers being fed by the transformer.
- Z is transformer percentage impedance [4], [5].

3.2.3. Single-Phase Distribution Transformer

Reticulation (see Figure 2)

This kind of reticulation involves three-phase and single phase medium voltage. Phases are tapped off the main three-phase line in pairs to single phase distribution transformers.

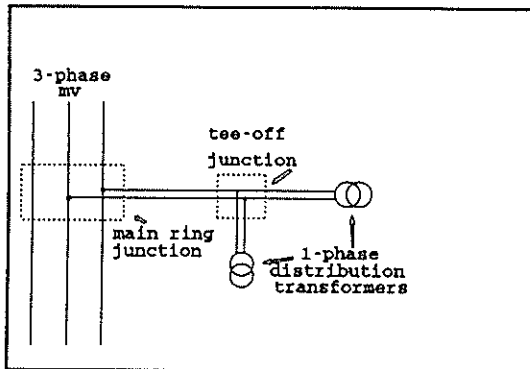


Figure 2: Single Phase Distribution Transformer Reticulation.

This situation has to be broken into two distinct areas where different equations are required to calculate current and volt drops:

- (i) The three-phase medium voltage main ring, from which single phase tap-offs are made.

Equations (1), (2), (3), (4), (5), (6) and (7) apply equally well to these junctions.

- (ii) Two core medium voltage tap-off's feeding single phase distribution transformers.

The per phase current in this MV feeder feeding this junction is given by:

$$I_{phase} = \frac{N * I_{max} * CF(N)}{a} \quad (8)$$

as we are dealing with single phase transformers.

The per phase volt drop in this MV cable is given by:

$$V_{phase} = \frac{2 * I_{phase} * \alpha_{mv} * L * UCF_s(N)}{1000} \quad (9)$$

as go and return paths have resistance, where :

- $UCF_s(N)$ is the appropriate unbalance correction factor for N consumers depending on the type of cable and/or transformer configuration.

The volt-drop within the transformer due to transformer impedance is taken into account using equations (3), (4), (5), (6) and (7).

3.2.4. Intermediate Voltage and Mixed

Reticulation Systems (see Figure 3)

The equations previously mentioned allow the designer to cope with intermediate voltage and mixed reticulation systems. Use of the correct equations will even apply to the design of single phase intermediate voltage reticulation systems.

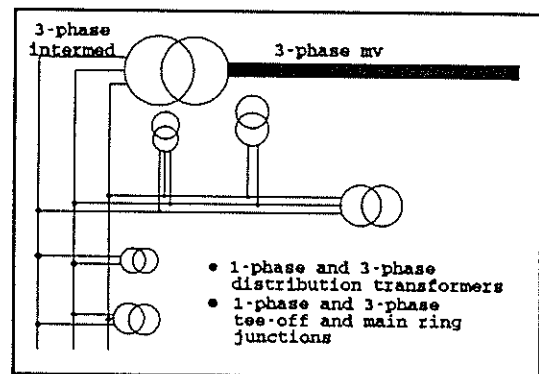


Figure 3: Mixed Reticulation Systems.

3.3. Medium and Low Voltage Optimisation

LV and MV is optimised as one system for all possible cable type combinations to give the required consumer termination voltage.

Consumer termination voltages are calculated from the following equations:

$$LV_{distrf} = [(MV_{supply}) - (\sum_1^x v_{drops}) - (MV_{droptrf})] * a \quad (10)$$

where:

- LV_{distrf} is voltage available at low voltage terminal of the distribution transformer.
- MV_{supply} is the main supply that the medium voltage is taken from.
- x is the number of cables that form a path from the main medium voltage supply point to the distribution transformer.
- v_{drops} are the voltage drops associated with the above-mentioned cables.

$$V_{consumer} = LV_{distrf} - \sum_1^y (v_{drops}) \quad (11)$$

where:

- $V_{consumer}$ is the voltage received at the consumer termination point.
- y is the number of low voltage feeders and service cables that form a path from the distribution transformer to the consumer termination point.
- v_{drops} are the voltage drops associated with the above-mentioned cables.

This paper does not cover LV calculations and assumes that LV volt drops and currents are available for the above calculation.

Performing iterations using equations (1) to (11) for all possible combinations of medium and low voltage cables, the most cost-effective design that meets consumer termination volt drop limits is found.

3.4. Other Thoughts on CART.

Development of CART does not stop after medium and low voltage optimisation. Further additions to the software may soon include: automatic cable routing and junction positioning; placement and costing of all electrical components, poles and pole furniture; advanced statistical algorithms for more accurate diversity determination; expert systems for population parameter assessment and expert help for reticulation design.

Medium voltage switching alternatives, system loss calculations and linking of CART to databases used by

HV and other MV analysis packages are just a few examples of what kind of improvements can still be made to the package.

4. CONCLUSIONS

Software implementation of the simple equations and method described in this paper provide an advanced design tool for a process that was previously tedious and time consuming. The added medium voltage design option allows the reticulation designer to experiment with the trade-off between a more complex medium voltage network and a simple low voltage network or vice versa. The improved CART gives the reticulation designer numerous design options and can be used to produce detailed electrification feasibility studies and final designs.

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NEW COMMANDS FOR CUSTOMIZING AUTOCAD IN AN ELECTRIC DRAWING ENVIRONMENT

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Customizing of any graphical software increases productivity and efficiency of CAD. This paper describes one way of customizing AutoCAD release 12 used for electrical drawings. It is achieved by user files, placed in separate directories so that the original software structure is not changed, thus enabling future AutoCAD upgrading. Particular review is given for new commands, created by lisp files. Each command is described through its function and illustrated on one of the figures.

1. INTRODUCTION

During the planning and design of a power network many drawings are produced which takes time. This time can be greatly reduced by making full use of a CAD system. Many of the graphic CAD packages are general-purpose drafting systems which can be adapted to meet a company's specific needs by customizing the software. This can be done by:

- implementation of own drawing standards
- creating symbol libraries
- automation repetitive tasks
- writing new user commands
- programming interface.

AutoCAD release 12 has been selected as a CAD system for the Port Elizabeth City Electrical Engineer's Department. It has been customized for all the various electrical drawing types in an effort to increase productivity and efficiency. The next paragraphs describe how it was done.

2. HARDWARE

AutoCAD release 12 has been installed on a NOVELL NETWARE 386, version 3.11 Local Area Network (LAN).

Each workstation comprises:

- Personal computer 486DX-50MHz, 256 kB RAM Cache, 16 MB RAM, 80 MB Hard disk (17 ms)
- Graphics Accelerator Card - Hercules Graphite
- Colour monitor - 21" Phillips Brilliance 2110
- Digitizer - 12" x 12" Calcomp model 33120
- Keyboard with built-in calculator.

and Colour plotter NovaJET 840.

3. FILE STRUCTURE

All files, used in drafting, can be divided as:

- original AutoCAD files
- user files.

Original AutoCAD files have their own structure, which is not discussed here.

User files, however, are files created by the user. They can be classified according to their function:

- program files:
 - menu files
 - lisp files
- drawing files:
 - symbol files
 - block files
 - job files.

It is important to store all user files correctly to achieve an effective working environment. They should be placed in separate directories to enable further AutoCAD upgrading without losing user files in the process. So, user lisp files are saved in separate CUSTOM directory, all menu files together with batch file and all slide libraries in separate DRAW directory (fig.1). This will ensure the original AutoCAD structure is not changed and there is an easy access to the user files.

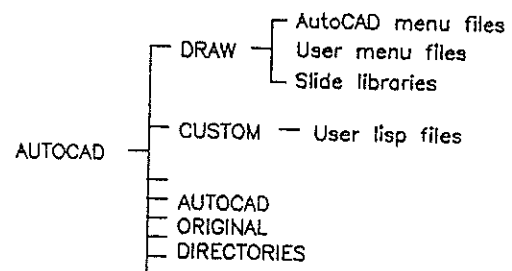


Figure 1: AutoCAD directory structure

All drawing files are placed under the same structured tree "drawings" (figure 2).

Job files consist of various drawings classified according to their type, as geographic (G), civil (X), architectural (W), electrical (Y), and mechanical (V) and placed in different subdirectories.

Each type is divided into its subtypes (see "Electric drawing" on figure 2).

All job files are created by symbols or/and blocks.

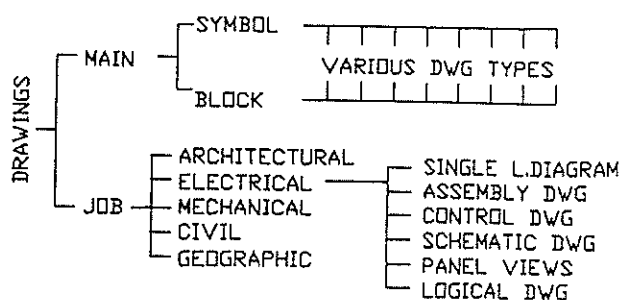


Figure 2: Drawing directory structure

Every drawing type has its own symbol library, so symbols are saved in a subdirectory according to its drawing type. This enables easy manipulation of every symbol group and automatic creation of icon symbol library via a scrip file.

Blocks are structured in the same way as symbols.

4. USER FILES

4.1 Menu files

A menu file is a text file containing AutoCAD command strings and macro syntax that defines the menu macro. It controls functionality and appearance of various menus and a pointing device. In addition to the original menu files the following are created:

USER.mnu
USER.mnx
USER.mnl.

The first one is a source file defining menus like: icon, pull-down, screen and tablet menu and a pointing device. It enables access to all symbol libraries and all user commands as well as AutoCAD commands.

The pull-down menu in USER.mnu has two new options: CUSTOM and SYMBOLS (figure 3).

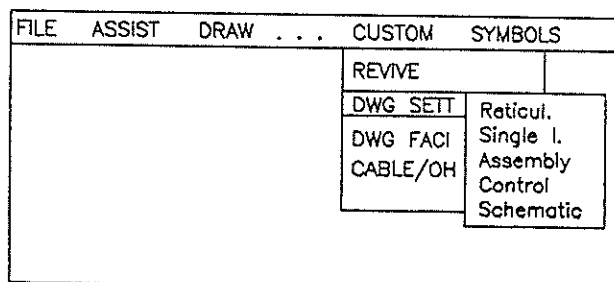


Figure 3: Pull-down menu with additional sections CUSTOM and SYMBOLS

All user created commands (paragraph 5) can be called from the "Custom" option while the "Symbols" option consists of various icon symbol libraries, which will be displayed on the screen for selection.

The menu file USER.mnx is a compiled source menu file.

USER.mnl is a menu lisp file which enables automatic loading, into memory, of all user commands that can be used by the "Custom" option. Users will, therefore, not need to load a lisp file (command) each time they need to use it.

4.2 Lisp files

These files are programs, stored in ASCII text format, defining user commands. They are loaded by USER.mnl file automatically if the USER menu file is active. Each file consists of a header, main routine and one or more subroutines with comments. The header gives an explanation about the use of the routine. The function of every command, developed by the user, is described in detail in paragraph 5.

4.3 Symbol files

All symbols are classified by two groups: - symbols
- blocks.

A symbol is a figure conventionally used on a drawing to represent an item.

A block is a part or assembly, created by symbols. It is a product of higher level standardizing.

While symbols represent basic symbols, blocks represent parts or assemblies like:

- standard substation in control diagram
- standard transformer bay in a single line diagram
- schematic diagram of overcurrent relay...

All symbols and blocks are drawn in a separate layer to control colour of where they occur. So they become the same colour as the current active layer colour. In that way the 11kV network will be green and 6.6kV blue etc., (figure 4).

Every symbol or block has a corresponding slide file. Slide files are used to create an icon symbol slide library, which will be displayed on the screen enabling symbol selection from the screen instead of the tablet. It is far easier and more efficient when working.

Some symbols/blocks consist of attributes (texts with special tag value) required for further drawing processing. Some attributes are visible on the screen and the others are not.

5. USER COMMANDS

These commands have been developed to improve work, release users from repetitive tasks, reduce time necessary for drafting and standardise some drafting functions like layers, texts, linetypes, etc. All of them are interactive and user-friendly and consists of validity checking.

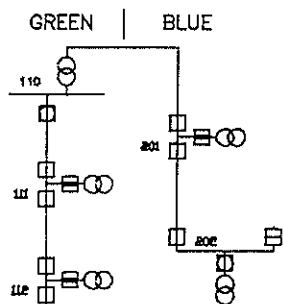


Figure 4: Control diagram
11/6.6 kV

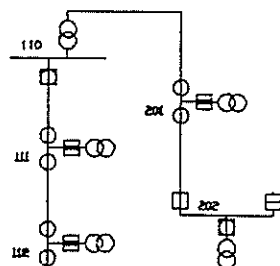


Figure 5: Symbol updating
by REVIVE command

5.1 "REVIVE" command

When changing an existing symbol, it is not applied automatically to an existing drawing in AutoCAD. The user can update the graphics of a symbol by use INSERT command specifying the full symbol path. This would require x of calling the "Insert" command and x of typing the full symbol paths for x of symbols updating. The old definition of symbol attributes will remain.

The new command "REVIVE" regenerates the drawing updating all existing occurrences of all used symbols. The drawing is automatically regenerated applying the new definitions of graphics and attributes of used symbols from one or more symbol libraries to all existing insertions of them. The User is only required to type REV or REVIVE and wait for it to end. Result of that is visible comparing figure 4 and 5.

5.2 Setting commands

Every drawing type has its own setting command. This is important for drawing standardisation to avoid different colours, linetypes, thicknesses or text heights for the same purpose. These commands define:

- standard layers with fixed colours and linetypes
 - standard text styles
 - standard line thicknesses
 - some drafting functions as grid, snap, linetype scale.
- Use of standard symbols and setting commands produces uniform drawing in the Department.

5.3 "OHLINK" command

Overhead line is shown as a line with poles in NRS 002 Proposed Amendment No.1. The line can be solid for existing or dashed for planned overhead line. Poles can be hatched (existing) or not (planned). AutoCAD does not cater for this line type with poles. The command "OHLINK" developed by the user, enables its drafting.

The User only needs to enter:

- type of overhead line (existing or planned)
- distance between two poles
- is it a line with or without stays.

It draws a line through the end points of the sections.

There are three options during the drafting:

End/Undo/<Next point>:

"End" means drafting of the last pole of the OH line. "Undo" erases the last drawn segment and can be repeated till the start point when all segments are undone. "Next point" continues to draw next segment of a line. The line will be drawn with stays in every corner if a line with stays is selected.

It is very important that all segments of an overhead line are the same line (entity) because of connecting the same characteristics from multiuser database. This command accepts the jointing of all segments into one polyline.

"OHLINK" consists of main routine and six subroutines. It checks user enter validity. It can be called as OH or OHLINK.

An existing and a planned overhead line are shown on figure 6, drawn using "OHLINK" command.

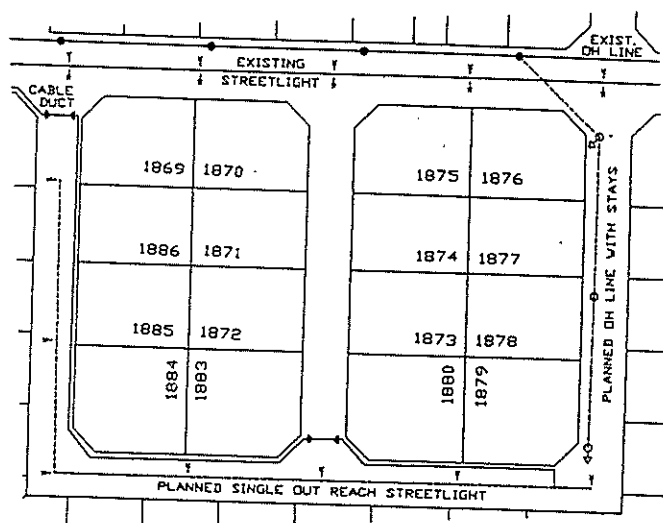


Figure 6: Reticulation drawing

5.4 Streetlight line

The command "STRLIGHT" enables drafting of streetlight line, existing or planned, with single or double out reach streetlight, cable or line combined with power overhead line. It can be called by typing in the full name STRLIGHT or short one SL from the keyboard or sectioning it from the "Custom" pull-down menu.

The User ought to enter:

- existing or planned streetlight
- single or double lights
- cable or overhead combined with power line
- distance between two poles.

There are three options when using this command:

End/Undo/<Next point>:

The "Next point" option draws the next segment of the line. The "Undo" option erases the last drawn segment while the "End" option draws last light finishing the command. All segments of a streetlight line are jointed together into the same polyline.

An existing cable double reach streetlight line and a planned single one are shown on figure 6.

5.5 "CSCALE" command

Sometimes the same symbol library is used, but with different drawing scales for the same symbol in various drawing types (reticulation drawings, substation layouts, etc.) to keep storage space to a minimum. On the other hand the user may want to draw part of a drawing in one scale (132 kV switchgear layout) and the rest in another scale (66 kV layout).

The new command "CSCALE" defines the drawing scale so that all symbols will be drawn in that scale until the user changes it. When inserting any symbol, the user no longer needs to worry about X or Y symbol scale. The 132 kV part of transformer bay layout is drawn in one scale while the 66 kV in the other scale (figure 7).

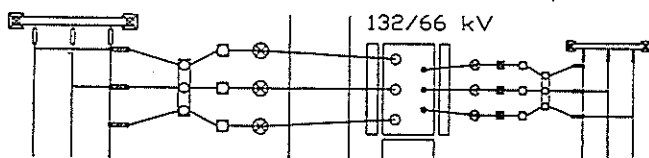


Figure 7: Transformer bay drawn by "CSCALE"

5.6 "DUCT" command

This command draws a cable duct of chosen length under any corner of a selected cable section.

The User will enter:

- type of duct (existing or planned)
- duct length
- start point
- angle to cable segment.

The command is interactive with validity checking. It can be called by typing DUCT or shortly DU. An existing cable duct is shown on figure 6.

5.7 "COUNT" command

The "COUNT" command counts the desired blocks printing their number as well as the last one used on the drawing. The User will enter a value of a string which is common for any particular block names and the computer will print:

Number of "YCA" blocks is: 6 The last one is: YCA006.

It is useful for automatic block naming in a drawing.

5.8 Extract of cables or overhead lines data

Cables or overhead lines are drawn as polylines or lines. They have associated strings which describe their characteristics. These strings are added by AutoCAD's "XDATA" command.

It would be necessary to extract these strings from the drawing for any further drawing processing.

This can be achieved by using the new command "LINKIT". This command will extract associated strings and write them to a separate text file. The User can call this command by typing LINKIT and then select objects whose strings are required to be extracted.

6. CONCLUSION

In most cases graphical software is a general-purpose drafting system. Tailoring it to meet our specific needs, means increase productivity and improve quality. It needs to be done very carefully so that the original software is not changed. This allows for any future upgrade of the software without undoing anything we have achieved.

Standardising of symbols and drafting rules should be applied from the very beginning.

The most repetitive tasks can be avoided by creating new commands which can achieve the same on our behalf. During standardising and customizing, future drawing processing should be taken into account.

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The uppermost level is for the control of the national transmission grid (132kV to 765kV), while the next level in the hierarchy deals with regional control (33 to 132 kV). SCADA

master stations are used on both levels of control.

The third level in the hierarchy is substation control via a micro-processor based RTU while the next level deals with reticulation control. RTUs mounted on poles have been developed to meet the requirements for the control of the reticulation network (33kV and below).

3 A NEW TELECONTROL PROTOCOL

If one looks closely at the control hierarchy in its entirety, it is clear that interconnectivity between all the devices comprising the system is a fundamental requirement if such a complex control structure is to operate correctly and efficiently. In order to realise this interconnectivity, a communications protocol was required which would provide the necessary functionality without compromising the real-time control performance requirements.

It was also necessary to consider the existing telecommunications infrastructure as only low baud rate lines were available. It was important to select the protocol so as to allow for upgradeability at a later stage. Further, it was essential that a decision regarding the protocol be made early in the implementation of the four level structure.

Due to the broad scope of the Four Level Control Vision, it was deemed unacceptable to commit to a proprietary protocol. In addition the international telecontrol protocol standardisation effort was a long way from producing final standards.

A strategic decision was made in 1987 to develop a telecontrol protocol specifically for Eskom's purposes, and the protocol was named Estel. It was recognised that the effort involved would be significant. However, in order to minimise the disadvantages of a user specific protocol, the following additional decisions were made:

- The development of Estel would follow the recommendations of the OSI model,
- The work of the IEC would be used as far as possible and all existing IEC

recommendations (at the time) would be adopted,

- Estel would eventually evolve towards full compliance with an internationally accepted telecontrol protocol standard.

4 ESTEL

Estel has evolved substantially since the first version was published and is currently on version 6.0 with features such as:

- Remote time synchronisation of slave stations.
- Transfer of bulk data to and from slave stations.
- Prioritised data retrieval functions and
- Special handling of overflow conditions on communication lines.

In order to select the most appropriate technologies for each particular component of the vision and to obtain the most cost-effective solution, Eskom applied a Structured Development Methodology (SDM) during the process of acquiring the telecontrol masters and RTUs. This consists of breaking a project into five phases, namely the functional specification, design, development, testing, commissioning and acceptance phases.

This process facilitated control over development projects and highlights potential problems at an early stage.

During the process of defining the Control Vision, Eskom identified a number of concerns regarding the implementation of a new protocol:

- The need to develop and test the implementation of the protocol against a standard was essential.
- Eskom required strict performance criteria to be adhered to for every SCADA master station of the Control Vision. This required a performance testing tool, since it was unlikely that sufficient RTUs would be available for testing.
- Maintenance of the equipment and protocol in the longer term.
- A high quality training mechanism was needed to be put in place to properly train staff.

The simulators and Bulk Data Generators were successfully used in all the implementations to test the functionality and performance of the protocol.

5 ESTEL IMPLEMENTATIONS

A sophisticated energy management system called ENCOR, was installed during 1991, which realised the first level of the Control Vision. This is a distributed system where the SCADA and energy management functions run on powerful back-end computers, with the interface to the energy network being managed by dedicated front-end computers. Estel has been implemented in the front end computers of the ENCOR system during 1992. This allows ENCOR to perform telecontrol at the main transmission substations where the new generation Enhanced Remote Terminal Units (ERTU) are installed.

The second level of the Control Vision consists of a system called SMART which is very similar in architecture to ENCOR with a well-defined path for upgradeability in terms of size and functionality. A contract has been placed for the acquisition of these systems and the first SMART was due to be commissioned during 1993. Since SMART and ENCOR are based on the same software architecture it reduced the cost and effort of implementing Estel on SMART.

Estel has also been implemented on the ERTU and 1993 saw the first time that a master, namely ENCOR, communicated to an ERTU in the field using the Estel protocol.

Finally, reticulation control has been made possible by the acquisition of Pole Mounted RTUs (PMRTUs).

The SMART master will also be required to communicate with the PMRTU central controller. This development is due to take place during the latter half of 1993 and will be a feature of the first SMART system to be installed.

6 ESTEL SUPPORT TOOLS

The requirement for Estel protocol simulators to provide a benchmark for the protocol was identified early in 1990. All development and

testing was to be performed using these products.

- The Estel Master Simulator facilitates the development, testing, commissioning and maintenance of slave devices. This simulator is capable of simulating the communications capabilities of a master station.
- The Estel Slave Simulator was designed to simulate the communications functionality of a slave station. The slave simulator can be used to simulate the communications of up to eight slave stations on one PC. This is useful when testing the multidropping of slave stations on a single physical communications link.

Eskom addressed the requirement for performance testing of SCADA master stations by means of a Bulk Data Generator (BDG) which is PC based Estel Slave system with a capability of simulating the communication of up to 480 RTUs. This system allows simulation of normal and bust-up conditions on the communication lines.

Maintenance issues have been addressed through the development of a product called the Estel Communications Line Monitor. The Line Monitor is used to passively monitor Estel communication lines. This device supplies raw and Estel decoded data in a user friendly environment.

The need for maintenance of the Estel protocol and related equipment was addressed by the development of the Estel Training package using computer based training software. This training medium removes the need for expensive class-room type training and the dependency on small numbers of protocol specialists for training.

7 IMPLEMENTATION ISSUES

The process of design, implementation, testing and commissioning the protocol on different products has provided Eskom with a number of insights into the process initiated in 1987. In particular the following points can be made:

- The decision to develop Estel is still very valid since standards for telecontrol protocols are not yet mature.

- The development of the support tools for testing the protocol simplified the management of the protocol implementations thereby reducing the overall risk.
- The development of the support tools addressed all the aspects of design, testing, maintenance and operation of the protocol. They will ensure the longevity of the protocol within Eskom.
- Through the involvement of the end-user during the design, specification and development of the products wide acceptance of the products and protocol has been achieved within Eskom.
- In addition, it is evident that user requirements have been addressed without sacrificing system performance or optimisation.
- By adhering to standards where they existed Eskom has simplified future upgrades to the protocol.

Problems experienced during the development of the protocol have resulted in valuable lessons:

- The implementations of the protocol in some instances created conflicting requirements in the protocol specification resulting in delays due to project schedules. However, often proper configuration management was implemented to resolve these problems.
- Certain aspects of the protocol could have been optimised to improve performance. However, these performance requirements of the protocol are not seen as a problem.
- The current implementation of the protocol is designed to operate on slow baud rate communication lines. However, Eskom has embarked on new communication facilities. This is due to Estel being a layered protocol.
- Eskom could have specified the interfaces between the layers of the protocol in much more detail. This would have shortened project schedules.

If circumstances at the time of choosing a protocol had been different then Eskom may have taken a different route. For example, if mature standard protocols that provided the required functionality had been available

Eskom has addressed current user requirements and allowed for an upgrade path

in the implementation of the protocol and related products.

8 PROTOCOL EVOLUTION

While Estel is generally accepted within Eskom, it is in Eskom's interests to ensure the acceptance of the protocol in the broader telecontrol community. The following steps have been taken to address this:

- Estel has been placed in the public domain, and the Estel protocol specification is freely available.
- The Simulator software is owned by Eskom and will be sold at a low cost.
- Telecontrol products from multiple suppliers are available that communicate using the Estel protocol.
- Estel will evolve to conform to an international standard telecontrol protocol.

With the implementation of Estel in the four level vision now a reality, the emphasis has shifted to ensuring the longevity of the protocol. This will be achieved by evolving Estel to meet new requirements and marketing the capabilities of Estel outside Eskom.

9 CONCLUSION

Eskom has made a strategic investment in the Estel protocol over the past six years. By resolving a number of concerns identified early on through sufficient protocol support tools and applying a structured development methodology, the implementation of the protocol has been achieved successfully.

If the reasons for committing Eskom to the Estel protocol are re-evaluated, the decision to develop Estel are still valid today.

Eskom is committed to the Estel protocol and is seeking ways to expand the user base and to ensure the continued applicability of Estel into the next century.

MODELLING, SIMULATION AND REALITY IN EDUCATION OF ELECTRICAL AND COMPUTER SCIENCE ENGINEERS

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Abstract

In the field of "Electrical Engineering" and "Process Automation" a concept has been developed for combined education in electrical engineering and applied computer science.

The classic analytical methods of mathematics are completed and replaced by numerical methods and the use of simulation software.

In co-operation with industrial firms a workplace has been developed for practical training and is presented at the conference.

About fifty experiments cover the fields of electrical engineering, measuring techniques, electric drives, automatic control, simulation.

1. Introduction

The development of computers and software has provided new tools for engineers. As a consequence new ways of methodological approach are to be developed in engineering education.

In theoretical education analytical mathematical methods have to be completed and replaced by simulation software.

In practical training computer aided measuring and control techniques have to be introduced.

2. Methodological Approach

The backbone of any solution of a technical problem is the knowledge of physical phenomena and their application. Reality has to be described by equivalent circuits,

models, set of algebraic or differential equations, arrays of measured data etc.

When using the computer, different to classic methods, first the technical problems have to be treated in an appropriate way to be solvable by this tool. The total problem has to be divided into modules, which can be worked out separately. Checked modules later can be used as submodules in greater systems.

3. Procedures

At the beginning of their studies students write small procedures. For example differentiation of measured data can be worked out. (Fig. 1)

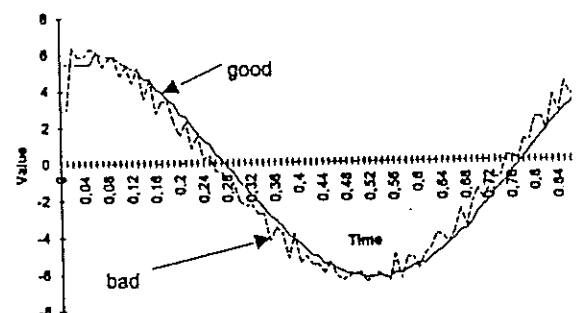


Fig.1: Differentiation of sine with superimposed noise

Differentiation is a dangerous procedure, because it can lead to numeric troubles. The number of calculated values and their accuracy have to be taken into account.

For the beginner it is an impressing experience, to learn, that the result becomes worse when the amount of data is increased.

3. Basic Electrical Engineering

Parameter Identification

Students have to test a coil with linked iron core. They have to consider the device as a "black box", measure transient voltage and current, find a model or equivalent circuit and determine the parameters.

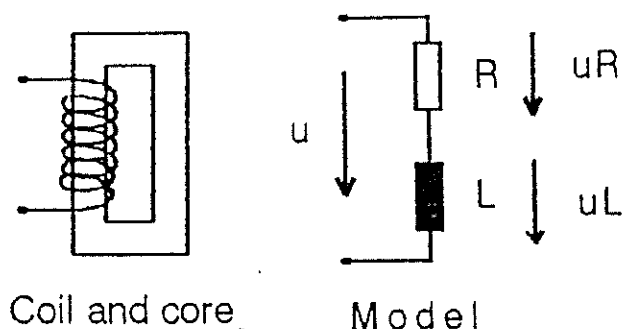


Fig.2: Reality and model

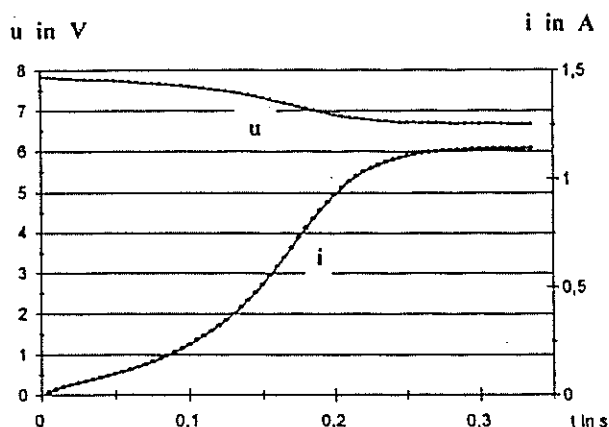


Fig.3: Transient voltage u and current i

The resistance R is calculated from the stationary values of voltage U and current I .

$$R = U / I. \quad (1)$$

During the transient process Kirchhoff's law is

$$u = u_R + u_L \quad (2)$$

or with ψ flux linkage

$$u = i * R + \frac{d\psi}{dt}. \quad (3)$$

With the definition of the differential inductance

$$L_d = \frac{d\psi}{di} \quad (4)$$

the equation becomes

$$u = i * R + L_d * \frac{di}{dt}. \quad (5)$$

Integration of the equation

$$\int u * dt = R * \int i * dt + \int d\psi * \frac{dt}{dt} \quad (6)$$

and

$$\psi = \int u * dt - R * \int i * dt \quad (7)$$

leads to the simulation model in Fig.4.

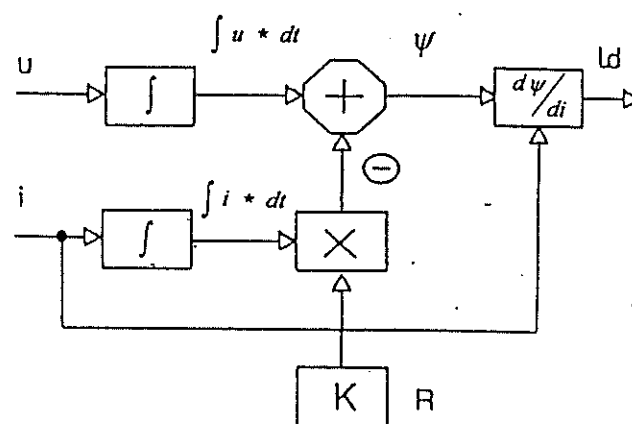


Fig.4: Basic Simulation model for calculation of differential inductance

Students are free to choose their software just as they like (Turbo Pascal, C, Xanalog, Excel etc.).

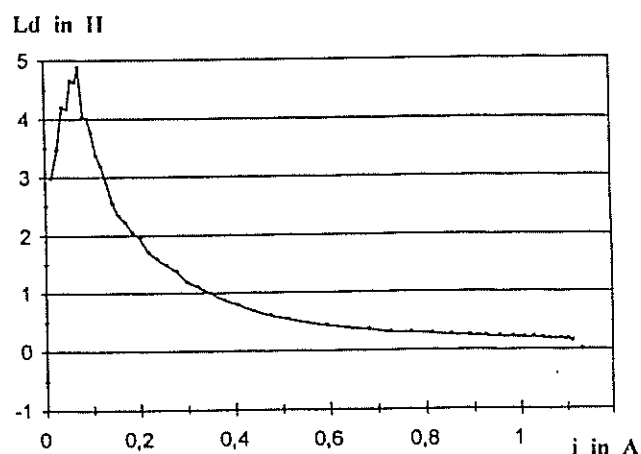


Fig.5: Differential inductance as function of current

4. Electric Drives

Stationary Characteristics of an Asynchronous Motor

This example deals with the automatic measurement of real stationary characteristics of asynchronous machines. To run the experiment the speed is adjusted following a schedule provided by a personal computer.

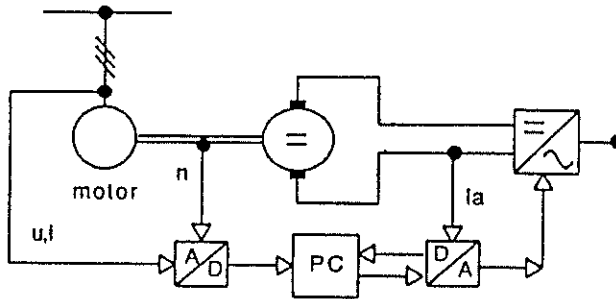


Fig.6: Automatic testing of motors

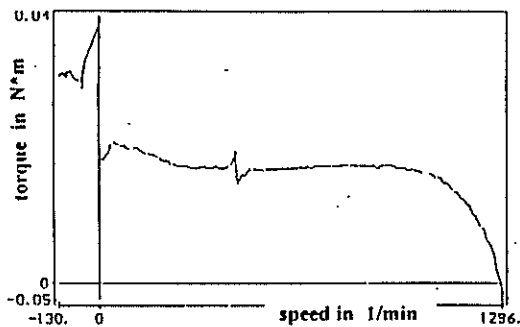


Fig.7: Stationary torque curve

Dynamic Torque Characteristic of an Asynchronous Motor

Due to transient phenomena in the magnetic field there is a strong oscillating torque during the starting period. To study this topic the speed is measured by means of a digital oscilloscope.

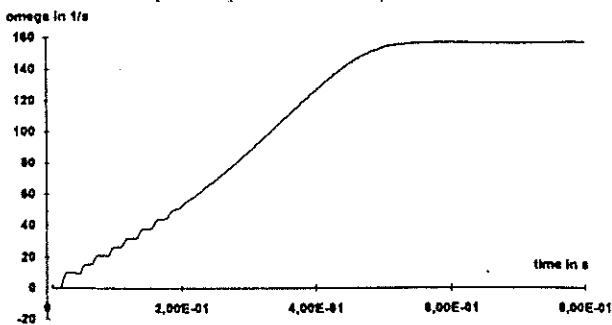


Fig.8: Starting of asynchronous motor, speed as a function of time

At the beginning of the start period there are irregularities in the curve. There is an oscillating component superimposed on the rising speed.

At no load the torque m can be calculated by the equation

$$m = J * \frac{d\omega}{dt} \quad (8)$$

with J ...moment of mass,

ω ...angular velocity

The appropriate processing of measured data depends on the intention of the experimenter. If he is interested in the overall acceleration of the mass or the average torque at any speed, he will smooth the measured data and then apply the differentiation procedure on the angular velocity (Fig. 9).

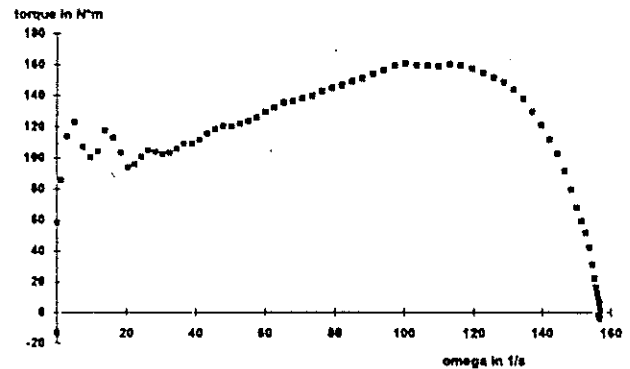


Fig.9: Dynamic torque curve

If he wants to know the pulsating torque during the starting period he will apply the differentiation procedure directly on the measured data.

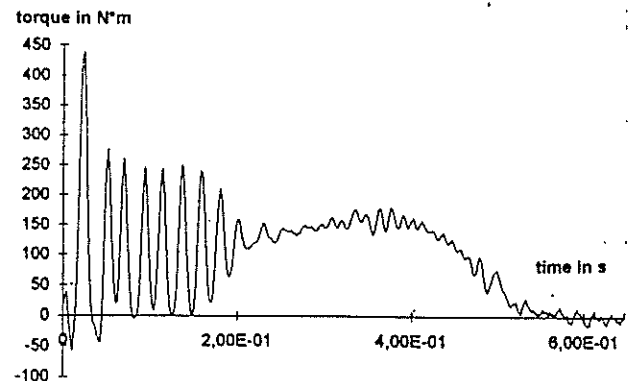


Fig.10: Starting of asynchronous motor, torque as function of time

5. Process Automation

Technical devices, which were studied in detail in the first years, are later combined to complex systems. As an example a power system consisting of diesel driven generators, battery, passive load and propulsion motor was to be simulated (Fig. 11).

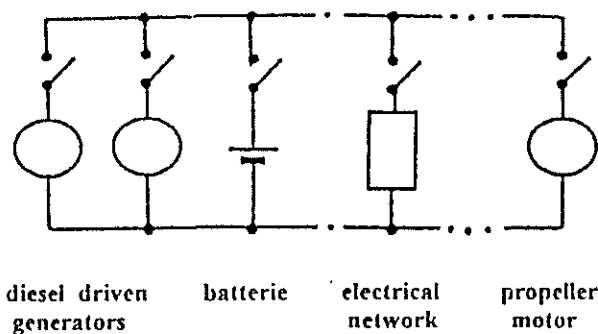


Fig.11: Power system

The simulation diagram is rather complicated (Fig.12). Each block represents a couple of subroutines in hierarchical structure. The results of simulation cannot be obtained by analytical methods.

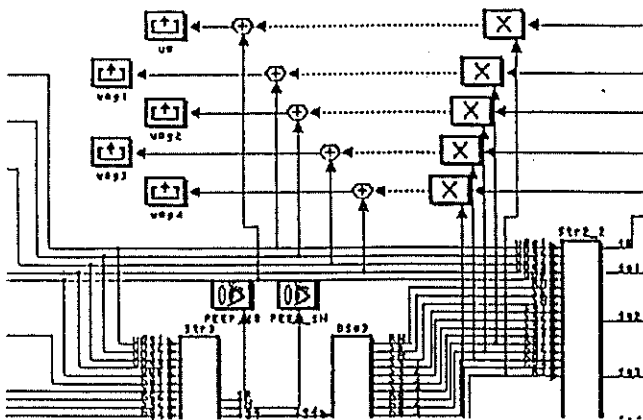


Fig.12: Part of simulation diagram "Power System"

8. Discussion

---The education is practice oriented. The used software tools and technical devices are the same as applied in industrial practice.

--Application of numerical mathematics is easier to understand than analytical theory.

--Emphasis therefore can be given to physical and engineering problems.

Basic difficulties arise in numerical mathematics when the method of approach is not appropriate to the physical problem. The unskilful application of a correct numerical algorithm to a correct equation may produce an incorrect result.

The proposed method of approach is applicable to many other subjects as mechanical engineering, control systems, biology, economics etc. (1).

9. Conclusion

Students should be educated from the very beginning of their studies to use the computer as a tool. Teaching of methodological approach therefore has to be changed to new methods.

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SELF EXCITED OSCILLATIONS IN POWER SYSTEMS AS AFFECTED BY EXCITATION AND GOVERNOR CONTROL

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Series capacitors are used to compensate the inductive reactance of long transmission lines, thus increasing their permissible straightaway transmission distance. However, improving the transfer capability of the transmission lines in this way may lead to resonance problems at subsynchronous frequencies. This paper deals with the analysis of self-excited oscillations in series capacitor compensated power systems.

The mathematical model developed includes turbine governors and results are presented to show their effects on the control of self-excited oscillations. The system performance over a wide range of excitation and governor parameter values are studied for the proper design of these controllers.

1. INTRODUCTION

Series capacitors have been extensively used in long-distance EHV transmission lines in many countries to improve the power system stability, but unfortunate experience of subsynchronous resonance (SSR) with generator shaft torques have demonstrated the need for caution in the use of such series compensation. Previous studies have shown that, under certain conditions, SSR can occur in synchronous machines connected through series capacitor compensated transmission lines either in the form of high frequency or low frequency self-excited oscillations.

Some work has been reported on the effect of various excitation control systems on the self-excited oscillations. Saito et al. [1] have demonstrated that a negative damping stabilizer, an excitation controller using the reactive power as a control signal, can effectively suppress high frequency self-excited oscillations. The effects of different types of conventional excitation regulators working on the proportional and derivative feedback signals of current, angle and voltage on both low and high frequency self-excited oscillations were investigated by Hurry [2]. The systems considered by the authors did not include the turbine generator torsional dynamics.

It has been reported [3,4] that speed governors have an adverse effect on the stability limits. This may be due to poor parameter values of the governor. The effect of an electrohydraulic governor on self-excited oscillations has been investigated in the past [4,5,6]. Such electrohydraulic governors with suitable gains

have been found to be effective in suppressing low frequency self-excited oscillations only. However, the simultaneous action of governor and excitation regulator was found to have an adverse effect on the systems stability. Moreover, in these studies, the authors have considered constant values of the governor parameters, thus the combined effect of excitation regulation and governor control on the self-excited oscillations has not been thoroughly investigated.

It is therefore important that the combined effect of governor action and excitation control on the self-excited oscillations be studied taking into account the torsional dynamics of the turbine-generator system. The aim of the paper is to suggest a systematic procedure for the simultaneous selection of the governor parameters of a power system which is equipped with forced action excitation control systems and of the controller parameters.

The present work is divided into three main parts. The problem of self-excited oscillations in an unregulated system is first examined using the Routh-Hurwitz method. Subsequently, the parameter plane technique is used to study the effect of excitation regulators on these self-excited oscillations. In the last section an electrohydraulic governor model is considered in order to investigate its effects on the power system in the presence of the excitation control schemes.

2. MATHEMATICAL MODEL

2.1 Power System Model

The small displacement model of the power system (Figure 1) representing a synchronous machine connected to an infinite busbar through a series-compensated long transmission line and provided with an excitation regulator working on a generalised feedback signal π is given by equation (1).

$$[A] \cdot [\Delta x] = 0 \quad (1)$$

where $[A]$ is a 4×4 matrix, and the elements A_{ij} 's are given in the Appendix.

$$\Delta x = [\Delta v_{fd} \Delta i_d \Delta i_q \Delta \delta]^T \quad (2)$$

The mathematical representation of the unregulated conventional synchronous machine is based on the well-known two-axis concept of Park's, and is therefore subject to the assumptions necessary for the derivation of these equations. Damper windings are

not included in the model. The transmission system is represented by a lumped line impedance in series with the compensating capacitor.

2.2 Torsional Resonance Model

In deriving the small displacement representation of the power system (equation (1)) the torsional mechanical modes of the shaft have not been accounted for. The effect of the shaft dynamics can be included by modifying the element A_{44} of the characteristic matrix of the power system (equation (1)) [3]. The simplified diagram of Figure 1 is used for the formulation of a lumped parameter model of a steam turbine-generator-shaft assembly.

2.3 System data

The studies have been made for a system with the following data:

$R_a = 0.005$ p.u.; $E_q/V = 1.0$; $X_d = 0.64$ p.u.
 $X_q = 0.64$ p.u.; $X_d' = 0.41$ p.u.; $X_e = 0.6$ p.u.;
 $T_r = 0.02$ s = 6.28 rad.; $H_G = 1884$ p.u.;
 $H_{T1} = 900$ p.u.; $D_{T1T1} = 0.6$ p.u.; $D_{T1G} = 0.6$ p.u.;
 $K_{T1G} = 22.5$; $T_{do} = 0.02$ s = 6.28 rad.

3. ANALYSIS

3.1 Unregulated System

The characteristic equation for the unregulated system is obtained by equating the principal determinant of the small displacement model to zero and is of the form:

$$a_0 p^7 + a_1 p^6 + a_2 p^5 + a_3 p^4 + a_4 p^3 + a_5 p^2 + a_6 p + a_7 = 0 \quad (3)$$

where the coefficients $a_0, \dots, a_7$ are functions of the system parameters.

Stability investigations are carried out using the Routh-Hurwitz criteria and results are provided in the form of boundaries separating the stable and unstable regions in the R/X_C plane.

Figure 2 shows the corresponding stability limit curves for the system. At low values of the system resistance, both high and low-frequency self-excited oscillations and torsional mode interaction are probable. Further analysis shows that the stability limit curve is very sensitive to D_{T1G} and D_{T1T1} . These studies provide an indication of the severity of the torsional interaction in power systems.

3.2 Regulated System

The effects of different excitation control schemes (current and angle regulators) on the self-excited oscillations are reported. The stability investigation is carried out by applying the parameter plane technique to the characteristic equation for the system obtained by equating the principal determinant of equation (1) to zero and is of the form:

$$a_0 p^{14} + a_1 p^{13} + a_2 p^{12} + a_3 p^{11} + a_4 p^{10} + a_5 p^9 + a_6 p^8 + a_7 p^7 + a_8 p^6 + a_9 p^5 + a_{10} p^4 + a_{11} p^3 + a_{12} p^2 + a_{13} p + a_{14} = 0 \quad (4)$$

where the coefficients $a_0, \dots, a_{14}$ are functions of the system and control parameters.

Analysis reveals that suppression of low frequency self-excited oscillations can be effectively handled by the use of suitable proportional current regulators (Figure 3). A common zone of stability (shaded area) is obtained for a wide range of load angle δ_L . The stability region is found to be larger than the corresponding one obtained through the use of the simpler model (no torsional dynamics). However, large values of both the proportional current regulator gain and the exciter time constant are needed. The incorporation of the derivatives of the current signal is found to extend the stable zone in the K_{1I}/K_{2I} plane. In the presence of the derivative current regulator, the common zone of stability is seen to be influenced by the larger value of the operating load angle as shown by the shaded area in Figure 4.

As regards high frequency oscillations, it is found that neither a proportional nor a current regulator can improve the situation. Likewise, it is found that angle regulators are also unable to extend the high frequency boundary.

Current regulators provide larger stable zones than the corresponding regions obtained by angle regulators in the vicinity of the low frequency boundary. The difference in the effect of these two regulators is quite significant when the torsional resonance model is used.

3.3 Simultaneous Excitation and Governor Control.

Figure 5 shows the effect of the proportional gain K_{OG} of an electrohydraulic governor and on the stability region in the T_r/K_{OI} plane. An increase in the value of K_{OG} considerably increases the stability region for practical load angle values. The addition of the first and second derivatives of the current signal is seen to extend the stable region. Furthermore, from Figure 6, it is observed that an increase in K_{OG} slightly increases the stable zone in the K_{2I}/K_{1I} plane.

The incorporation of the acceleration and the rate of change of the acceleration signals in the governor loop is observed to extend the stable zone. Such an improvement can also be achieved through the use of the acceleration feedback signal only (Figure 7). Figures 8-10 show that an increase in either the reheat coefficient or the generator inertia constant increases the size of the stable region for practical values of K_{OG} and T_r . Besides a decrease in either K_r or T_{tu} increases the stable zone.

Further analysis shows that the simplification made in assuming the governor to be directly attached to the generator and not to the turbine leads to optimistic results for the range of $K_{2G} < 2 \times 10^4$.

4. CONCLUSIONS

From these investigations it is clear that the shaft torsional dynamics should not be neglected. Effective application of the parameter plane technique has been

demonstrated to study the combined effect of excitation regulators in the presence of governors.

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6. NOMENCLATURE

i_d, i_q	d- and q- axis components of stator current
v_{fd}	field voltage
v_d, v_q	d- and q- axis components of voltage
X_d', X_q'	transient reactance in d- and q-axis circuits
Ψ_d, Ψ_q	d- and q- axis stator flux linkage
R_a	armature resistance
X_d, X_q	d- and q-axis synchronous impedances
$X_d(p), X_q(p)$	d- and q-axis operational impedances
	q-axis operational impedance
$X_{d\Sigma} = X_d' + X_e$	
$G(p)$	field operational function
p	differential operator d/dt
Δ	prefix to denote small changes about the initial operating point
o	suffix to denote small changes about the initial operating point
R_e, X_e	external resistance, external reactance
X_C	series compensating capacitor reactance
X_{ad}	d-axis magnetizing reactance
R_f	field winding resistance

T_{do}	open-circuit field time constant
T_e, T_r	exciter and rectifier time constants
K_r	reheat coefficient
T_{rh}	turbine reheat time constant
T_{tu}	turbine time constant
E_q/V	level of excitation
$K_{0\pi}, K_{1\pi}, K_{2\pi}$	proportional, first and second derivative gain of generalised feedback signal
$g_D(p)$	governor transfer function
$X_{d\Sigma} = X_d + X_e; X_{q\Sigma} = X_q + X_e$	
$R_\Sigma = R_a + R_e; G(p) = (X_{ad}/R_f) / (pT_{do} + 1)$	
$X_{d\Sigma}(p) = (pT_{do}X_{d\Sigma}' + X_{d\Sigma}) / (pT_{do} + 1)$	
$X_{d\Sigma}' = X_d' + X_e$	
λ	degree of damping $p = -\lambda + jw$
w	angular frequency
H	inertia constant (no torsional dynamics)
H_g, H_{T1}	Turbine-generator inertia constants
K_{T1g}	Turbine-generator shaft stiffness
T_1, T_2, T_3, T_4	governor constants

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8. APPENDIX

Elements A_{ij} 's of the small displacement model in equation (1):

$$\begin{aligned}
 A_{11} &= 1 & A_{12} &= -K_\pi(p) M(p) \\
 A_{13} &= -K_\pi(p) N(p) & A_{14} &= -K_\pi(p) L(p) \\
 A_{21} &= (1-p^2)G(p) \\
 A_{22} &= p^2 X_d(p) - X_d(p) + pR_\Sigma + X_C \\
 A_{23} &= -(2pX_d(p) + R_\Sigma) \\
 A_{24} &= p^2 \Psi_{qo} + 2p\Psi_{do} + v_{do} - p i_{qo} R_\Sigma \\
 A_{31} &= -2pG(p) & A_{32} &= (2pX_d(p) + R_\Sigma) \\
 A_{33} &= p^2 X_q(p) - X_q(p) + pR_\Sigma + X_C \\
 A_{34} &= -p^2 \Psi_{do} + 2p\Psi_{qo} + v_{qo} + p i_{do} R_\Sigma \\
 A_{41} &= i_{qo} G(p) & A_{42} &= -(\Psi_{qo} + i_{qo} X_d(p)) \\
 A_{43} &= \Psi_{do} + i_{do} X_q(p) \\
 A_{44} &= H p^2 = H p^2 + g_D(p) \text{ (modified to include torsional dynamics)} \\
 K_\pi(p) &= [K_{0\pi} + pK_{1\pi} + p^2 K_{2\pi}] / [pT_r + 1)(pT_e + 1)] \\
 H' &= H G p^2 + (D_{GG} + D_{TIG}) p + K_{T1G} - (D_{T1G} p + K_{T1G})^2 / [H_{T1} T_2 p^2 + (D_{T1T1} + D_{T1G}) p + K_{T1G}] \\
 g_D(p) &= [K_{0g} + pK_{1g} + p^2 K_{2g}] [pT_4 K_r + 1)(pT_2 + 1)] / [pT_1 + 1)(pT_3 + 1)(pT_4 + 1)]
 \end{aligned}$$

For a current regulator :

$$M(p) = i_{do}/i_o, N(p) = i_{qo}/i_o, L(p) = 0,$$

For the angle regulator :

$$M(p) = 0, N(p) = 0, L(p) = 1$$

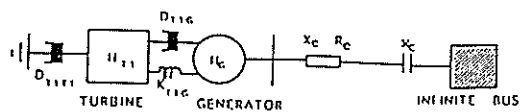


Figure 1: System investigated.

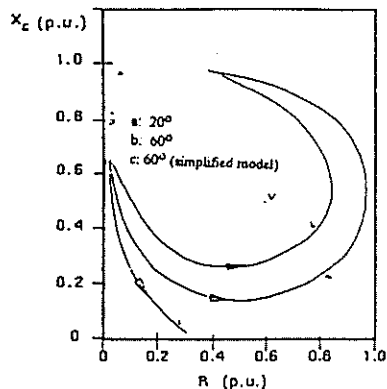


Figure 2: Stability limit curves (torsional dynamics).

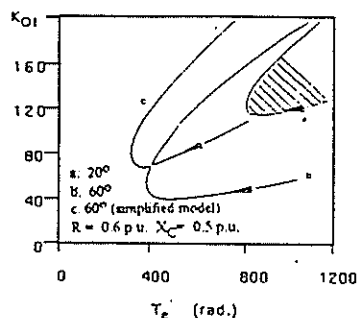


Figure 3: Stability zones (no governor).

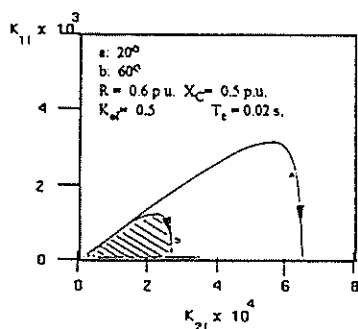


Figure 4: Effect of derivative current regulator gains.

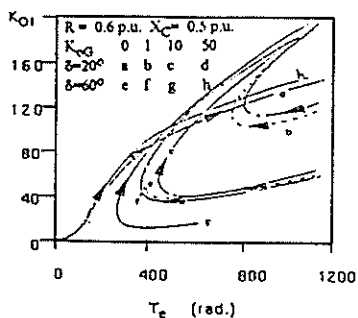


Figure 5: Effect of proportional governor gain.

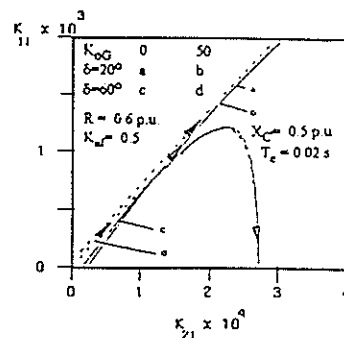


Figure 6: Effect of governor/current regulator gains.

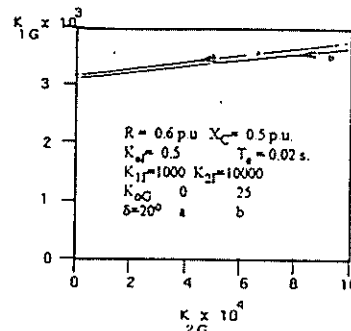
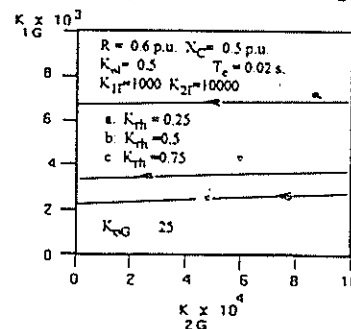
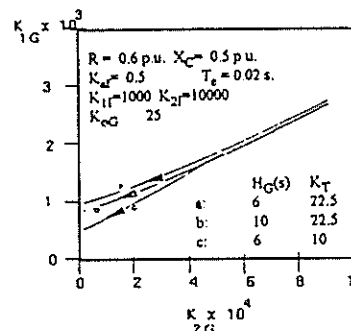
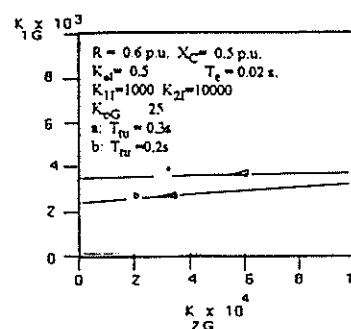


Figure 7: Effect of governor derivative gains.

Figure 8: Effect of K_{th} on the stability zone.Figure 9: Effect of T_{tu} on the stability zone.Figure 10: Effect of H_G and K_T on the stability zone.

COORDINATION OF GENERATION PLANNING IN THERMAL POWER SYSTEMS

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ABSTRACT

This paper outlines a practical coordination approach of long- and short-term optimization. The approach uses pseudo prices as coupling variables. To take uncertainty aspects into account, fuzzy set theory is used to evaluate the pseudo prices. In the paper, various aspects of the proposed approach are discussed, and its application to a real utility system is illustrated.

1. INTRODUCTION

A very important task of electrical utilities is the economically optimal operation of the generation units concerning the primary energy allocation. Based on the load prediction and the specific properties of a utility the optimal operation schedule has to be determined. A lot of constraints have to be considered during the optimization algorithm. It is possible to formulate the optimization problem by using Mixed Integer Programming, Dynamic Programming or decomposition methods. Although a formulation of the optimization problem seems to be rather simple, the listed methods are not applicable for solving the problem in a one-step procedure. The requirements concerning computation time and storage cannot be satisfied for large scale power systems on available computers.

The complete optimization problem is divided into several steps [1] as shown in Figure 1.

The electric utilities often have to take long-term energy constraints into account. These constraints are

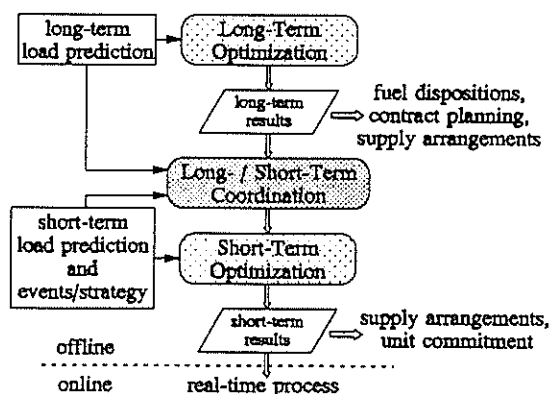


Figure 1: Optimization steps

caused by contractual limitations (e.g. take-or-pay-contracts) or by technical or administrative reasons (e.g. due to emission control). Corresponding to the time horizon of the long-term constraints the long-term optimization normally concerns periods up to several years [2]. The model accuracy of the long-term optimization cannot be very high in order to reduce the number of variables. The achieved results of the long-term optimization are energy allocations for the power plants.

The short-term optimization deals with unit commitment decisions. Here the short-term operational constraints e.g. minimum up- and down-times, power ramp rates, start-up costs, balance of load and generation and spinning reserve requirements have to be satisfied. The time horizon of the short-term optimization is up to one week.

1.1 COORDINATION OF LONG- AND SHORT-TERM OPTIMIZATION

Long-term energy constraints cannot be incorporated into the short-term optimization algorithm directly because the constraints are related to the long-term time horizon. Nevertheless it is necessary to take the long-term energy constraints into account indirectly by using pseudo fuel prices as coupling variables [3] or by forced allocation of energy values for the units [4], [5].

Numerical investigations [6] have shown the difficulties of finding an effective and automatic coordination method between long- and short-term optimization which is applicable under different operational conditions.

Figure 2 shows the sequences and combined action of long- and short-term optimization starting at a time T_0 . Based on the long-term optimization at T_0 , every week a short-term optimization cycle is done until the deviations between the optimization results and the real generation process exceed prespecified limits (here assumed after 4 weeks). Then a new long-term optimization cycle with updated system data is needed as new basis for following short-term optimizations.

In power systems uncertainties concerning load

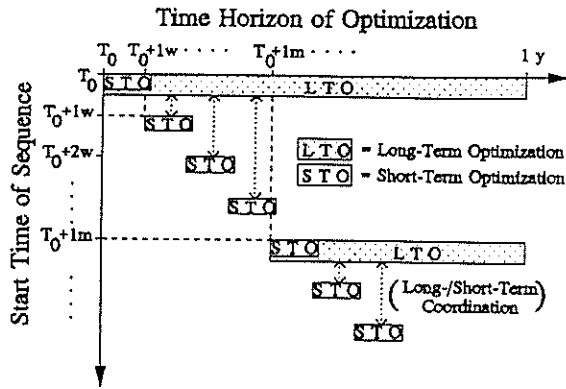


Figure 2: Sequence and combined action of long- and short-term optimization

prediction and real consumption or unexpected unit breakdowns occur. Significant changes cause new optimization runs with updated data. Small deviations have to be suspended by a robust optimization procedure. Such an often occurring deviation is the difference in the energy amount of the long-term and the more accurate short-term load prediction and the real consumption during the short-term time horizon.

An energy allocation approach for the long-short-term coordination can fail if the energy amount of the long-term load prediction differs from the short-term load prediction [6]. Therefore the proposed coordination algorithm is based on the use of pseudo fuel prices.

Another request for the coordination method is the assistance of variant investigation. Different strategies have to be compared in order to select the best one for the actual scenario. An easy description of the different strategies by linguistic terms is useful. A mathematical tool for formulating uncertainty and provide the use of linguistic terms is the fuzzy set theory. In section 3 a fuzzy set approach for the coordination problem taking uncertainties and linguistic terms into account is described.

2. MULTI-CRITERIA COORDINATION PROBLEM

The investigation of different variants is often related to the investigation of different strategies (for example, a strategy, keeping the satisfaction of the energy constraints always on the secure side, or in another case, a cost optimal strategy taking the risk of potential constraint violation). The investigation of different strategies can be represented by a multi-criteria decision problem as shown in Figure 3.

Every unit U_i , $i=1, \dots, m$ has to be assessed by a rating r_{ij} with respect to each criterion C_j , $j=1, \dots, n$. The height of the value r_{ij} reflects the accomplishment of

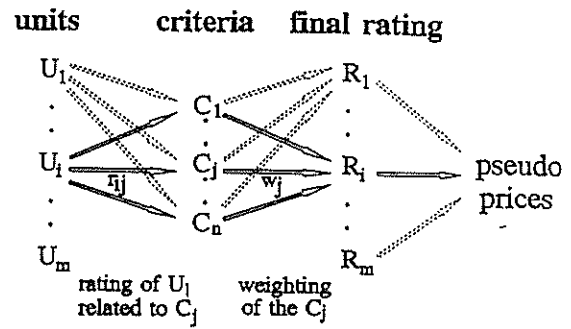


Figure 3: Multi-criteria decision problem

criterion C_j for unit U_i . The importance of each criterion C_j is modelled by a weighting coefficient w_j . Then unit U_i receives the weighted average rating or final rating

$$\bar{r}_i = \frac{\sum_{j=1}^n w_j r_{ij}}{\sum_{j=1}^n w_j} \quad (1)$$

The pseudo prices will be derived from the final ratings $\bar{r}_i$, $i=1, \dots, m$ as described in section 3.

A set consisting of six criteria is chosen to represent the above described problem in mathematical terms.

I. unit utilisation factor during the short-term period

$$r_{i,I} = \frac{E_{i,\pi}^{long-term}}{P_i^{max} T_\pi} \quad (2)$$

II. specific fuel costs of a unit at bestpoint operation

$$r_{i,II} = \frac{\max_{k \in (N)} \{cost_k(P_k^{max})\} - cost_i(P_i^{max})}{\max_{k \in (N)} \{cost_k(P_k^{max})\} - \min_{k \in (N)} \{cost_k(P_k^{max})\}} \quad (3)$$

III. possible violation of lower energy bounds of a unit

$$r_{i,III} = \max \left\{ 0, \min \left\{ 1, 1 - \frac{E_{i,lt} - E_{i,lt}^{min}}{\delta_i} \right\} \right\} \quad (4)$$

IV. possible violation of upper energy bounds of a unit

$$r_{i,IV} = \max \left\{ 0, \min \left\{ 1, 1 - \frac{E_{i,lt} - E_{i,lt}^{max} - \delta_i}{\delta_i} \right\} \right\} \quad (5)$$

V. possible violation of lower energy bounds of unit groups

$$r_{i \in \{L\}, V} = \max \left\{ 0, \min \left\{ 1, 1 - \frac{E_{i,lt} - E_{i,lt}^{\min}}{\delta_i} \right\} \right\} \quad (6)$$

VI. possible violation of upper energy bounds of unit groups

$$r_{i \in \{L\}, VI} = \max \left\{ 0, \min \left\{ 1, 1 - \frac{E_{i,lt} - E_{i,lt}^{\max} - \delta_i}{\delta_i} \right\} \right\} \quad (7)$$

In (4)...(7) critical surroundings of the lower and upper energy limits of unit i resp. of set $\{L\}$ of units are indicated by δ_i resp. δ_L . Using these equations, all the ratings assume values in the interval $[0,1]$. The ratings denote discrete values not reflecting the uncertainty of data. In the next section the application of fuzzy set theory to the coordination problem is described.

3. USE OF FUZZY SET THEORY

It is possible to represent uncertainty by the notion of fuzziness [7], [8], [9]. A fuzzy set is indicated as $S=(X, \mu_S)$, where X is the definition space of the fuzzy set, and $\mu_S(x)$, $x \in X$ the membership function of the set. The membership functions considered in the following sections can be characterized by the properties:

- normalized, bounded, convex with finite support
- piecewise continuously differentiable
- monotonic non-decreasing in the range $[-\infty, c]$ and monotonic non-increasing in the range $[c, \infty]$, where c is called core value with $0 \leq c \leq 1$ and $\mu(c)=1$.

Instead of discrete values r_{ij} and w_j now fuzzy sets S_{Rij} and S_{Wj} are used. Fuzzy sets allow to model possibility distributions similar to probability distribution in statistics. In Figure 4 membership functions for the rating and weighting factors are shown. The core value c of a of the membership function μ_{Rij} is given by the discrete value computed from the corresponding formula (2)...(7). Not only the core value c but also numerical values within an interval around the core value are members of the fuzzy set. The membership of the numerical values decreases with growing distance to the core value c , while the membership of the core values is 1. The width of the interval is a measure for the uncertainty of the rating.

Considering the different criteria it is obvious, that some ratings can be determined rather certain, others are uncertain. The specific fuel costs of a unit are e.g. fixed values whereas the utilization factors and the violation of energy constraints are uncertain because they depend on the uncertainty of the load prediction. In this approach the uncertain ratings are generally modelled as intervals of 5% around the core values.

A formulation of the weighting factors in linguistic terms for example: "unimportant", "moderately important", and "important" is provided by the fuzzy set theory as shown in Figure 4. The results presented in section 4 have been evaluated using 5 equidistant fuzzy weightings as a compromise between accuracy and lucidity in the number of different weightings.

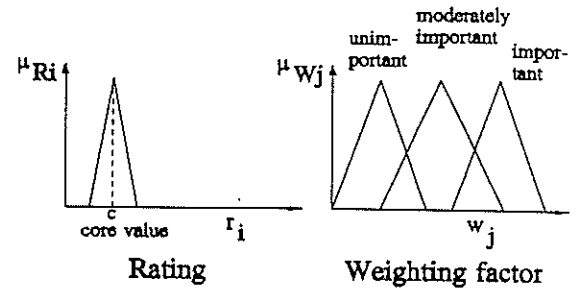


Figure 4: Membership functions of ratings and weighting factors

Considering the multi-criteria problem described by (1), it is possible to apply the theorems given in [7], in order to evaluate the fuzzy set of the final rating $S_{\bar{R}i}$.

The pseudo prices could be derived from the fuzzy sets $S_{\bar{R}i}$ by determining the centre of gravity of the integral below the according membership function $\mu_{\bar{R}i}$. Using these pseudo fuel prices the short term optimization yield the results presented in the next section.

4. NUMERICAL RESULTS

The short-term optimization is done by a program based on Lagrangian Relaxation [5]. In this program start-up costs, minimum up- and down-times, power ramp rates for the units as well as power balance and spinning reserve constraints are taken into account. The solution procedure is applied to a real system with an installed maximum generation of ca. 7000 MW.

Table 1 contains information about the fuel, the unit type and the corresponding energy constraints. Three different long-term energy constraints occur. The gas units have to satisfy a take-or-pay constraint called "ng". The own coal-fired units are supplied by domestic coal "dc" and imported coal "ic" which is much more cheaper than the domestic coal. Based on an agreement between the coal-mining industry, the utilities and the government for the domestic coal a minimum amount has to be consumed, although it is more expensive than the imported coal. Caused by contractual dispositions the imported coal is limited upside. A combined energy constraint for domestic and imported coal can be formulated called "dcl/c". For the

Table 1: generation system data

fuel	type	no.	energy constraint
nuclear	own unit	1	-
gas	own unit ("ng")	6	take-or-pay
gas/ hard coal	own unit ("ng/dc/ic")	1	take-or-pay
hard coal	own unit ("dc/ic")	4	take-or-pay
	contract ("ct")	11	take-or-pay
others	own unit	3	-
	contract	4	-

contracts a common long-term energy constraint called "ct" exists.

The following examples present short-term optimization results achieved with on the proposed coordination algorithm for a week during the winter. Based on the long-term optimization it is assumed, that any long-term energy constraint except that one called "ct" can be satisfied without problems. Those contractual units which are involved in observance of "ct" are expensive compared to the others. The combination of the economical point of view and the fact of an energy limitation of "ct" yields the optimal utilization of these contracts at their lower energy limit.

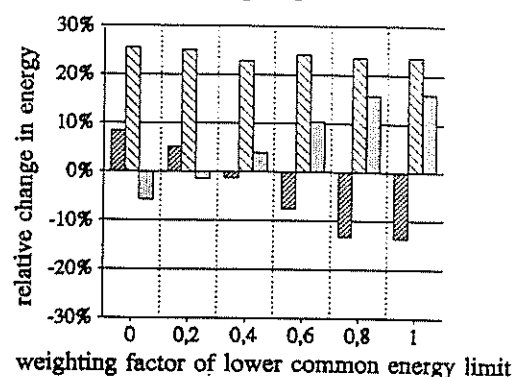
A different energy amount of the load compared to the load prediction can easily affect the quality of optimization results. If the real energy amount is lower than predicted, the constraint might be not observed. If the real energy amount is higher than predicted, the utilization increase of the contracts might be not economical.

Two different load scenarios are investigated:

- Short-term load prediction with a 3% higher peak load as in the long-term prediction ($\Rightarrow$ 3.2% higher energy amount)
- Short-term load prediction with a 10% lower load valley as in the long-term prediction ($\Rightarrow$ 2.4% less energy amount)

In Figure 6 the achieved results of the short-term variants for the scenarios a), b), and c) are shown. The energy consumption in dependence on the weighting of criterion V (see section 2) is presented. Five different weightings with equidistant core values in the range

a) computation with higher peak load



b) computation with lower load valley

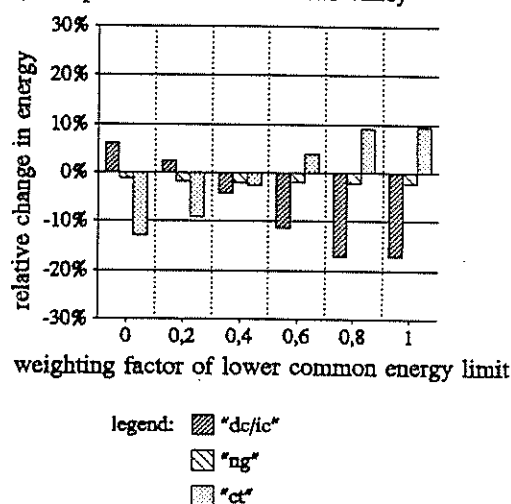


Figure 5: Relative change of energy consumption in dependence of the weighting of criteria V

from 0.0 to 1.0 have been investigated. Because only "ct" is critical utilized at the lower limit, the energy consumption of "ct" is directly influenced by the weighting of criterion V. Increase of the energy consumption of "ct" causes automatically a decrease of energy consumption of another primary energy. In this case this interchange takes place between "ct" and "dc/ic" because the gas units are peak load units, which cannot be less utilized without violating the reserve constraints. While in scenario a) the energy consumption of "ct" is for weighting factor 0.2 less and for 0.4 higher than in the long-term optimization, it is necessary to chose weightings of 0.4-0.6 in order to achieve a similar energy consumption of "ct" in the case of lower load (scenario b)). The planning engineer can easily investigate different strategies (for example: secure observance of criterion V, or risky economic optimal operation) by using different weightings of criterion V.

Analog effects of dependency between the weightings of other criteria and the influence on the according criteria could be noticed. Specific strategies can be formulated by sets of weightings of criteria I - V.

5. CONCLUSIONS

A new fuel allocation method has been proposed where pseudo fuel prices are used as coupling variables. Uncertainty of the long-term optimization results is taken into account by using fuzzy set theory to evaluate the pseudo prices.

Different strategies concerning the fuel allocation can be easily studied by weighting the criteria. It is possible to formulate linguistic descriptions of the criteria weighting so that investigations related to different strategies could be done.

The computation time of the coordination algorithm is about 2 seconds CPU-Time on a VAX-Workstation 3100.

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NOTATIONS

i	unit index
j	criteria index
$\{N\}$	set of units
U_i	i-th unit
C_j	j-th aspect
r_{ij}	Rating of i-th unit relating to j-th aspect
w_j	weighting of j-th aspect
$\bar{r}_i$	final rating of i-th unit
$E_{i,st}^{long-term}$	energy of i-th unit allocated from long-term optimization during short-term period
P_i^{max}	maximum power output of i-th unit
T_{st}	short-term period
$cost_i$	specific fuel costs of i-th unit
$E_{i,lt}$	long-term energy consumption of i-th unit
$E_{i,lt}^{max}$	upper limit of long-term energy consumption of i-th unit
$E_{i,lt}^{min}$	lower limit of long-term energy consumption of i-th unit
δ_i	value characterizing the critical area around the energy limits
$\{L\}$	group of units with a common energy constraint
μ_S	membership function of fuzzy set S

STABILITY OF A GENERATOR LINKED TO A POWER SYSTEM THROUGH A WEAK INTERCONNECTION

M C Coetzee
(Eskom)

This paper addresses the two major different states of power system stability, namely transient and dynamic. It deals with some of the network conditions that can develop into any of these states. Furthermore it will deal with possible solutions that can be implemented in a power system to solve potential stability problems.

1.0 INTRODUCTION

The generation of electrical power in a large transmission network is by synchronous machines. One problem in transmitting the power from the generating units to the load centres is for all the generators, connected to the network, to continue rotating in synchronism. The concept of power system stability is the ability of generators to maintain synchronism and to return to a steady state operating point following a system disturbance.

There are two major aspects of stability that can be associated with a power system, namely transient and dynamic. The main difference between these states is the time frames as indicated below.

Transient stability occurs in the time period of 0 - 1 second. The main components that will play a role during this time period will be the inertia, duration of the fault and the strength of the power system.

Dynamic stability occurs in the time period of 1 second to 1 minute. The main components that will play a role will be the strength of the power system, automatic voltage regulators (AVRs) and governors of each machine.

2.0 OVERVIEW OF STABILITY

2.1 TRANSIENT STABILITY

Transient instability is characterised by a remotely located power station that is connected to the network via insufficient transmission lines, thus forcing the power through a power bottle neck.

Transient instability occurs when the machine is subjected to a large impact, usually a fault. The fault will generally be sustained for a short time period during which the terminal voltage will be reduced and the ability to transmit power will be affected.

The following equations are the basic equations that relate to transient stability:

$$P_e = \frac{V_1 V_\infty}{X_e} \sin \delta \quad (1)$$

$$P_m - P_e = M \frac{d^2 \delta}{dt^2} \quad (2)$$

where P_m is the mechanical power output of the turbine

M is $2 \cdot H$ and H is the inertia constant of the rotor of the generating unit.

A disturbance can be subdivided into three stages, namely prefault, fault and post fault. During the prefault condition, the network will be intact.

During the fault time period the terminal voltage will be reduced and the power generated P_m cannot be transmitted due to insufficient transmission capacity and therefore the excess energy will be stored in the rotor and the rotor will accelerate as long as the fault is effective according to equation (2). The generator does not generally go unstable in this time period. To clear the fault, one of the transmission lines will be removed and therefore the external reactance increases.

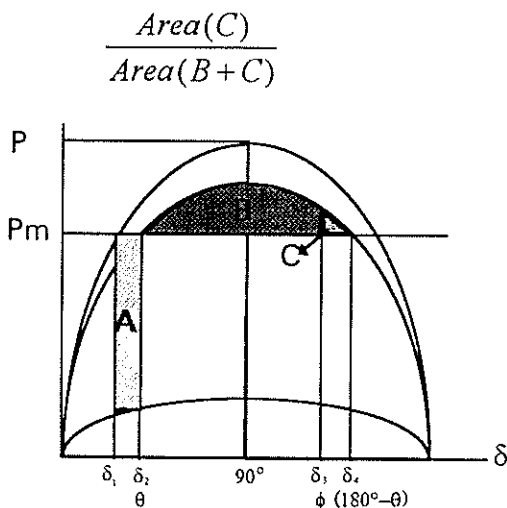
During the fault, the generator speed increases causing the angle δ , which is an integral of speed, to also increase from δ_1 to δ_2 when the fault is cleared, as indicated in figure 1. After the fault is cleared the power transferred increases according to equation (1). The generator continuous speeding up until δ is sufficient for the power transmitted in the electrical system to be equal to P_m . At this point the generator starts slowing down although the angle continuous increasing. The angle increases until such time that the energy associated with area A equals that associated with area B as indicated in figure 1.

$$\begin{aligned} \text{Energy} &= \int P d\delta \\ &= \int \left(\frac{P}{\omega}\right) d\delta \end{aligned}$$

Since the change in rotor speed is very small for periods of concern, $\text{energy} \propto \int P d\delta$.

Theoretically for stability to be maintained the energy stored in the rotor during the fault (area A) must equal or be less than the energy transmitted through the transmission system (area B). This stability margin is defined as the ratio of the total possible energy available (area C) divide by the actual energy required (area B).

Therefore the stability margin can be defined as follows from figure 1:



Area C

$$\begin{aligned} &= \int_{\phi}^{180^\circ - \theta} P(\sin \delta) d\delta - P_m(\pi - \theta - \phi) \\ &= P(\cos \theta + \cos \phi) - P_m(\pi - \theta - \phi) \end{aligned} \quad (3)$$

Area B + area C

$$\begin{aligned} &= \int_{\theta}^{180^\circ - \theta} P(\sin \delta) d\delta - P_m(\pi - \theta - \theta) \\ &= 2P \cos \theta - P_m(\pi - 2\theta) \end{aligned} \quad (4)$$

The stability margin is:

$$\begin{aligned} &= \frac{P(\cos \theta + \cos \phi) - P_m(\pi - \theta - \phi)}{2P \cos \theta - P_m(\pi - 2\theta)} \\ &= \frac{P/[P_m(\cos \theta + \cos \phi)] - (\pi - \theta - \phi)}{2P/[P_m \cos \theta] - (\pi - 2\theta)} \end{aligned} \quad (5)$$

It should be noted that while the generator is speeding up the damper windings behave as an induction generator and effectively reduce area A and while the generator is slowing down they act as an induction motor effectively and reducing area B.

The following factors are considered from a transient stability aspect when planning the transmission system.

The implementation of a fast acting AVR. The AVR does increase the voltage behind X' and if it is fast acting can have an effect in spite of the large time constants involved.

The protection and breaker clearing time on all the feeders out of a power station must be fast. The longer it takes to clear a fault, the more energy will be absorbed in the generator rotor.

Sufficient transmission lines must be built out of a power station to ensure that X_e will not increase to a too large value on the outage of a line.

If the above cannot be implemented various other methods can be considered such as:

- series compensation in lines.
- braking resistors at the power station.
- fast valving on the power station turbines.
- generator dropping
- limiting the power output of the power station.

2.1.1 Practical Application Of The Stability Margin

Eskom experienced a Transient Stability problem at one of its power plants. This power plant is connected to the system via long transmission lines. Electrically it is far from any other power plants and therefore with an outage of a line or during fault conditions the external reactance increases drastically. To solve this problem in the short term, from an operational point of view, the stability margin is determined and the output of the station reduced to meet an expectable margin. Eskom is planning to install a generator dropping scheme (Transient stability protection scheme) to solve this problem in the long term.

2.2 DYNAMIC STABILITY

Dynamic stability is the ability of a machine in the system to adjust to small load changes without losing synchronism. An immediate power increase is required if we assume a load change that is large enough to be recognised by a machine. Sudden step changes in power to the turbines are not possible and therefore the power increase will come from the generator rotor rotational energy until the governors adjust the power input to the various generators. During this time period the change in machine voltage, speed and current will be different for each machine in the system depending on the unit size, design and electrical location in the system. Therefore each machine will contribute to the load increase depending on the external impedance that each machine will see. Each machine has its own natural frequency of response and will oscillate until damping forces damp these oscillations.

Figure 2 is a representation of a single machine connected to an infinite bus which was developed by the authors in reference 2. It is a useful model to study dynamic oscillations. It also shows the coupling of the electrical and mechanical systems which can result in small generator oscillations about a steady state operating point. The feedback network outside the dotted box can be real or imaginary and may add or subtract from the damping term D . When the net damping is negative, oscillations may originate in a power system that appears to be in a secure steady state condition. Any normal load fluctuation or minor switching of power equipment on the power system may initiate these oscillations which will increase in time.

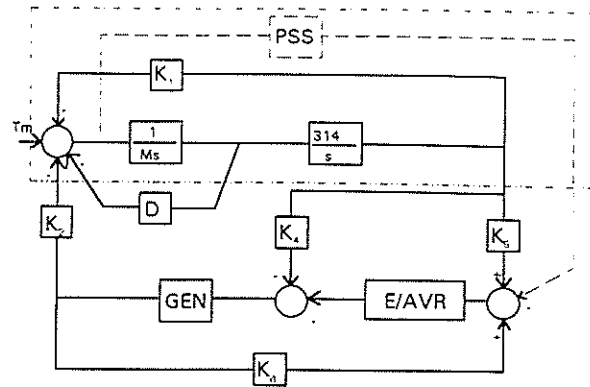


Figure 2: Single machine connected to infinite bus.

The characteristic dynamic behaviour of a machine can be expressed by the swing equation:

$$\frac{2H}{Ws} * \frac{d^2\delta}{dt^2} + \frac{D}{Ws} * \frac{d\delta}{dt} + K_1\Delta\delta = 0 \quad (9)$$

To illustrate when oscillations can be expected, assume $\delta = Ae^{mt}$

$$\text{then } m = \frac{-\frac{D}{Ws} \pm \sqrt{\left(\frac{D}{Ws}\right)^2 - \frac{8HK_1}{Ws}}}{\frac{4H}{Ws^2}}$$

$$\text{If } \frac{D^2}{(Ws)^2} < \frac{8HK_1}{Ws} \text{ then oscillations will occur.}$$

The natural frequency of oscillation of a machine is:

$$\omega_n \approx \frac{\sqrt{K_1Ws}}{2H} \text{ radians/second}$$

The term $K_1\Delta\delta$ is the action of the transmission system and acts to accelerate or decelerate the rotor to the steady state operating point. This term is called the synchronising power. The stronger the transmission system, the larger K_1 would be and therefore providing more synchronising power to the generator. K_1 is calculated as follows:

$$K_1 = \frac{dP_E}{d\delta} \Big|_{\delta_0} = \frac{E_q' E_0}{X'd + X_e} \cos \delta_0 \quad (7)$$

A factor that can aggravate dynamic stability is a fast acting automatic voltage regulator (AVR). The fast AVR is good for transient stability but introduces negative damping into the generators, therefore introducing oscillations in a frequency band of 0.2Hz to 2.5Hz. The natural damping of the system is represented by term D in equation (6) and will damp any oscillations unless a larger source of negative

damping is introduced. Such a source can be contributed by the normal feedback control of an AVR. To cater for the AVR action, the swing equation (6) must be extended to include an extra term $K_2 \Delta E' q$.

$$\frac{2H}{W_s} * \frac{d^2 \delta}{dt^2} + \frac{D}{W_s} * \frac{d\delta}{dt} + K_1 \Delta \delta + K_2 \Delta E' q = 0 \quad (8)$$

As it can be seen from figure 4 the term $K_2 \Delta E' q$ will be added to the synchronising power, $K_1 \Delta \delta$. The stability of a machine is depended on the value of this term. The imaginary part of $K_2 \Delta E' q$ will be added to the damping term ($\frac{D}{W_s}$) and the real part of $K_2 \Delta E' q$ will be added to the synchronising power and must be > then zero. The value of this term can be influenced by the strength of the transmission system or the terminal voltage.

For a small load change in the system, a power station will swing with respect to the rest of the system and this swing results in an oscillation frequency. This is called a local mode oscillation because the behaviour is localised to that power station. When two or more power stations are swinging with respect to the rest of the system an inter-area mode of oscillation will occur.

Solutions for dynamic stability are to take the generator off AVR, retune the AVR or add a power system stabiliser (PSS). A PSS, acting through the voltage regulation, can provide significant damping of dynamic oscillations if it is tuned and installed properly. Various input signals can be used as input to the PSS, namely the change in accelerating power, speed or angle.

2.2.1 Practical Application In Eskom For Dynamic Stability

Undamped oscillations were detected at one of Eskom's pump storage schemes in the pumping mode. The pump storage scheme is connection to the system via two long transmission lines, whenever one of these transmission lines is out, the external reactance is very high. The generators are dynamically stable in the steady state but if any switching occur in the network, undamped oscillations develop and increase with time. The exciters at this plant are very fast acting and therefore aggravate the situation. Eskom solved this problem by reducing the gain of the AVR so that it would respond slowly to any disturbance in the network. This is only a short term solution and a

PSS will be installed to solve the dynamic stability problem of this plant at a permanent basis.

3.0 CONCLUSIONS

Stability problems occur in weak transmission systems. Transient stability happens before operators can act. Transmission systems must be designed to ensure it is transiently stable at all times. Dynamic instability generally can be rectified by very fast operator action.

Dynamic instability can occur after a large disturbance and it is necessary to consider both transient and dynamic stability when studying large disturbances.

There are solutions that can be implemented in the short and long term for either of these stability states.

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APPRAISAL OF NIGERIA'S FIRST EXPERIENCE AT LARGE SCALE STEAM TURBINE POWER GENERATION

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ABSTRACT

The paper appraises the 720Mw Sapele Station, Nigeria's first large scale steam turbine power generating plant. The study period is from 1978 to 1992. The plant's availability, utilisation and average capability factors based on both the installed and available capacities were low. Within an average of six years, four of the six sets were put out of operation by turbine blade and boiler failures. Forced outages were mostly caused by external disturbances that were preventable. Comprehensive planning before the establishment of such stations by developing countries is advocated.

1.0 INTRODUCTION

The first large scale electric power generating station built in Nigeria comprised of four gas turbine (GT) sets, each capable of generating about 10Mw. It was in 1963. In spite of the fact that the economy of large scale favours the establishment of steam turbine (ST) plants, that did not happen until fifteen years after. In between, a two-set GT plant of installed capacity of 72Mw and a hydro plant which had four kaplan type machines with installed capacity of 320Mw were commissioned respectively in 1966 and 1969. The strategy was to make the maintenance crew of the National Electric Power Authority (NEPA) move from acquainting themselves with the easier-to-maintain GTs to the hydro sets before contending with the conventional but most-difficult-to-manage STs. The first ST plant was therefore established in 1978 in Sapele.

From the forgoing, it would appear that a good groundwork was made before the station was built and as a result an equally good performance would be expected. This paper proceeds from a study to appraise how the plant had fared from the time the first set was commissioned in September

1978 to the end of 1992. It is expected that findings from this study will be useful in planning for future steam plant development.

2.0 THE SAPELE ST STATION

Sapele is a town in the Delta State of the country. It was chosen for the building of the ST station because of the large reserve of natural gas round the area, (the gas was being flared at the time), the acceptable bearing foundation of the ground and the presence of sufficient cooling water for the condenser and bearing oil provided by the nearby Ethiope River. Preliminary work was started in 1974 and the first unit was commissioned in September, 1978 and the last in October, 1980.

The installed capacity of the station is 720Mw comprising of 6 x 120Mw units. Each unit has a two cylinder turbine - a single flow high pressure (H.P.) cylinder and a double flow low pressure (L.P.) one. The sets which are of the non-reheat condensing type have five stages of regenerative feed heating with the turbine shaft rotating at 3000 revolutions per minutes (rpm). Other relevant technical data are as follows: Steam flow rate 455,000 kg/hr, Steam temperature at turbine stop-valve 510°C and Steam pressure at turbine stop valve 87 bar.

3.0 PLANT PERFORMANCE

ST 1: ST 1 was commissioned on 12th September, 1978 as the first unit and it was forced out in June 1984 as a result of a number of problems. The set had its first major problem in September, 1979, and it was off for a month. It was involved in a fire disaster in June, 1980 and was not fully repaired until August, 1981. Early in 1984, refractory failures around the burner boxes gave rise to tube leakages. The unit was shut down in February and April for repairs. Burner and air supply ratio adjustments were

made to reduce boiler vibration.

High vibration was also noticed occasionally on turbine No. 2 bearing and it was therefore, necessary to do alignment adjustment. This did not solve the problem and investigation revealed that turbine No. 1 bearing was overheated and worn out. It was also discovered that there was a problem of blade failure. The set was therefore marked out for rehabilitation.

ST 2: This unit was commissioned on 1st February, 1979 but was forced out in June, 1987 as a result of LP turbine blade failure. ST 2 had its first problem, though relatively minor, when it was out of operation in April, 1980 for 20 days. It however developed major problems towards the end of 1983. It was back in operation on 5th February, 1984, for only six days before being forced out again for 183 hours. It was again shut-down on 1st April, 1984, and investigation revealed the following problems, among others: generator coolers and condenser tube leakages, turbine LP and HP blade failures, appreciable silica deposits on both HP and LP blades and nozzles. This unit continued tottering in performance even after it was overhauled in 1985.

ST 3: ST 3 started operation on 22nd March, 1979, but had been on shut-down since February, 1989. It was out for a total of 66 days in 1980, and 22 days in 1981. Early in 1984, it underwent repairs of furnace walls and tube leakage at windbox for some burners. Investigation of excessive turbine eccentricity revealed excessive silica build-up and HP blade failure. The unit was therefore out of service from 12th March, 1984 for the rest of the year. In March, 1985, it was recommissioned after it had been overhauled. However, in May, 1988, investigation of the high bearing turbine vibration revealed some wear on the last row of the LP turbine blades. This was the problem that led to its shut-down.

ST 4: This unit is one of the two that were still in service by the end of 1992. It was commissioned on 2nd November, 1979. In 1980, and 1981, it was out of service for a total of about 60 days. The boiler was overhauled in 1984. This was followed by repairs of leaking boiler tube and HP No. 4 and 5 feed heaters. Continuous excessive turbine vibration led to the derating of the output to 90Mw on suspected silica build-up. It was forced out in July, 1990 due to unit transformer bus duct ground fault caused by ingress of rain water through the cable duct. The unit returned to service in March, 1991, and has continued, though this has been on an on-and

-off basis.

ST 5: ST 5 was commissioned on 24th April, 1980, but was forced out in March, 1986, as a result of a fire incident on the unit transformer. In 1984, suspected silica deposits necessitated the derating of the unit to 90Mw output. The set had been constantly bugged by various problems such as faulty control oil system, boiler refractory failure, boiler tube leakages and boiler feed pump defects.

ST 6: This set is the second one of the units that were still in operation by the end of 1992. It was commissioned on 19th August, 1980. Between March and May, 1984, it was down for 1460 hours due to leakage and safety valve problem. The boiler and turbine were overhauled but the unit could not attain the 120Mw rating because of control oil system instability and attemperator control equipment problem. The furnace flooding problem which came after the boiler overhaul became worse in 1988. The set was subsequently derated officially to a maximum of 100Mw in 1990, when poor vacuum caused by fouled condenser tube was added to its problems.

Table 1 shows the performance levels of the sets. Availability factor is the ratio of number of hours a unit has been available for power production to the total number of hours in the period considered; average capability factor is the ratio of the average generating capability to the installed capacity; and utilization factor is the ratio of energy generated (Mwh) to the installed capacity of the unit in Mwh for the period.

4.0 SURVEY OF TRIPS

Trips were mostly of two types - those that originated as a result of faults in a particular machine and those that were caused by system disturbances in the grid. While the former affected the particular unit concerned, the latter resulted in the total collapse of the station.

In 1988, 1989, 1990 and 1991, there were 23, 14, 26 and 10 system disturbance induced trips. There seems to be no trend. The details of trips occasioned by internal problems in ST 4 and ST 6 (the only operational units) are given in Table 2. It is known that while problems with other components may be easily fixed, relatively, those of the turbine, especially the blades, usually put the set off for longer periods. Findings summarized in Table 2 confirm this.

5.0 MAINTENANCE HISTORY

Overhaul of the units is recommended after every 16,000 hours. This maintenance requirement has, however, not been followed as data in Table 3 show.

6.0 OBSERVATION AND DISCUSSION

An analysis of the observed performance of the units shows that :-

- * All the sets have low levels of performance as was pointed out in section 2, 3, and 4. All the six sets have been derated, mostly by as much as 25% after operating for as little as 30,000 hours or less than six years.

- * Incidence of turbine blade failure is high, given the relatively young age of the units. Most of this seems to come from silica deposits which suggests that the water/steam purification system is not adequate. It is known that silica is soluble in steam and its solubility increases rapidly at temperatures above 260°C(1).

- * The number of trips is high and this is not good for the health of the units. There seems not to be enough protection of the station from external disturbances.

- * Maintenance history is poor. The non-availability of spare parts and dearth of experienced staff seem to be the major drawbacks in the pursuit of an effective maintenance programme.

7.0 CONCLUSION

To reduce the incidence of trips caused by external disturbances, the installation of under-frequency relays at load centers, to isolate the station, (without necessary tripping the units) is suggested. Because of the number of complicated auxiliary equipment and the consequent high requirements of spare/accessories and the relative difficulty in handling steam compared with air and water (the other common working fluids) it is suggested that countries that have problems with hard currency sourcing, and which do not have personnel with adequate training and experience, should ensure that comprehensive programmes are in place to tackle those issues before establishing large steam turbine stations.

It has been stated that in the Nigerian environment and circumstance, gas turbines are more viable than steam turbines (2), however, from the view point of the economy of large scale, steam turbines are favoured. It may therefore be necessary if steam plants are finally chosen, to start with smaller units which are less complicated to run.

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TABLE 1: PERFORMANCE LEVELS OF UNITS AND PLANT

UNIT/ PLANT	AVAILABILITY FACTOR		UTILISATION FACTOR		AVERAGE CAPABILITY FACTOR
	INSTALLED CAPACITY	AVAILABLE CAPACITY	INSTALLED CAPACITY	AVAILABLE CAPACITY	
ST 1	0.17	0.40	0.26	0.10	0.59
ST 2	0.39	0.65	0.47	0.29	0.74
ST 3	0.42	0.56	0.47	0.36	0.86
ST 4	0.49	0.49	0.36	0.36	0.73
ST 5	0.32	0.68	0.47	0.22	0.69
ST 6	0.66	0.66	0.50	0.50	0.76
PLANT	0.41	0.57	0.42	0.31	0.73

TABLE 2: YEARLY AVERAGE TRIPS FROM INTERNAL PROBLEMS (1990 - 1992)

MAJOR COMPONENT WHERE PROBLEM WAS LOCATED	ST 4	ST 6
Boiler/Furnace	7	7
Condenser	2	3
Turbine	1	-
Feed Pumps/Feed Heaters	5	2
Generators	2	4
Total	17	16

TABLE 3: MAINTENANCE HISTORY OF UNITS

UNITS	No. OF OVERHAULS	DATE OF LAST OVERHAUL	HOURS RUN BEFORE OVERHAUL	HOURS RUN SINCE LAST OVERHAUL	HOURS SINCE COMMISS- IONING	THEOR- ETICAL No. OF OVERH- AULS
1	NIL	-	-	-	20,409	1
2	1	APRIL 1985	34,708	12,839	47,547	APPROX 3
3	1	MARCH 1985	24,045	24,757	48,802	3
4	1 (ONLY BOILER)	JUNE 1984	18,547	30,966	49,513	3
5	NIL	-	-	-	35,258	2
6	1	DEC. 1984	22,605	41,843	64,448	4

ENHANCING THE ELECTRICAL ENGINEERING COURSE AT CURTIN UNIVERSITY

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This paper describes some initiatives taken to enhance the image of the electrical power course at Curtin University. By revising the format of the course and publicising some of the activities of the department the image of the course has been enhanced and student numbers increased.

1. INTRODUCTION

At Curtin University, the School of Electrical and Computer Engineering offers four year engineering courses in the areas of Electrical Power, Communications and Computer Engineering. The number of final year students in the school is approximately sixty and the division of students between departments is currently about equal. This equitable situation is the result of a number of initiatives which have been undertaken in recent years. Prior to these initiatives the numbers in the power stream had declined alarmingly and the viability of the course was becoming questionable. (An experience shared by other universities in Australia, [1])

The course structure at Curtin is shown in Fig 1. Students proceed from a common engineering semester to three semesters of general electrical engineering and then specialise for the final two years.

1	Common first semester in engineering		
2	Common study program		
3	for students in the School of		
4	Electrical and Computer Engineering		
5	Computer	Electrical	Electronic
6	Engineering	Engineering	Engineering
7	specialised	specialised	specialised
8	units	units	units

The electrical power course with its emphasis on heavy current applications is quite distinct from other courses in the electrical engineering domain. The major areas within the course are :

Machines
Power Electronics
Control
Power Systems

While mention of these topics may bring a glow of pride to academics in departments of electrical engineering they do not excite similar feelings amongst second year students. These students are confronted with the choice of three courses and in the past most have found either computer engineering or communications more attractive than the power course.

2. STRATEGIES IMPLEMENTED

To increase the number of students opting for the power course (and ensure the survival of the department) various strategies were adapted.

(i) The experience with machines at the second year level was made more rewarding.

(ii) Some of the exciting things happening in the final years of the power course and at post graduate level were brought to the attention of students.

(iii) The units in the final years were revised to make them more modern in content and appearance.

2.1 SECOND YEAR UNITS

In the common second year, students previously studied two introductory machines units accompanied by traditional laboratory work. The laboratory program was a reflection of the facilities available in the machines lab., a collection of ac and dc machines brought out to terminals on benches. The configuration has hardly changed in the last fifteen years. Students were required to perform traditional lab tests, as described in the lab manual, to determine operating characteristics, machine parameters etc. From these lab sessions, most students gained the impression that

Fig. 1 Course structure at Curtin University

electrical power is a static, unfashionable discipline firmly rooted in the past. This impression was reinforced by the lecture program which also had changed little in the last decade.

The new approach we have adapted with the lectures is to broaden the syllabus by introducing some of the modern developments in the power field and reserve the in depth treatment later years. At this level; an appreciation of power concepts is felt to be more important than a detailed understanding of particular machines. This concept of a light first pass over a subject followed by successive forays at greater depth - (for those students who are still interested), is quoted by J R Greene in an interesting article which appeared recently in the IEEE. [2]. The philosophy existing at Curtin seemed to be, "since most students will not be choosing the power course let's try to teach them all about power while we still have them".

In the laboratories there have also been significant changes. The traditional format of lecture followed by laboratory has been replaced by a structure which completely separates lectures and laboratories. Laboratory units have become independent entities with their own learning agenda. Thus in second year, students enrol for a unit, Electrical Laboratory, which consists of: five weeks of power labs followed by five weeks of computer labs followed by five weeks of electronics labs. The structure of this new lab program lends itself to a more 'student driven' approach than in the past. Topics for investigation, with library references, are given as 'lab prelims' and students come to the lab with circuits and test procedures that they have sourced for themselves and consider appropriate. Of course at this level some guidance is still required and two staff are available in the lab to give advice, check wiring etc.

One of the lab sessions is to be spent in the School's 486 lab, running a power circuit simulation. The new generation of simulation programs such as PSpice and Simulink, which include schematic capture, make the simulation process relatively easy to implement. It is important that students see a connection between computers and the power discipline.

This revised format for the laboratory programs is continued in the later years of the course. [3]

2.2 PUBLICISING ACTIVITIES

At Curtin one of the features of the course is that during the final three semesters students spend considerable time doing a project. Some of the projects are undertaken in conjunction with local industry, others are associated with the research work going on

in the department. [4]. While there is some debate regarding the educational merit of undergraduate projects, [5] students in general enjoy doing project work and often achieve impressive results. This year we have introduced project seminars in our department, whereby final year students are required to give an oral presentation of their work to staff and other students. It is intended that second year students form part of the audience and will be enthused by some of the work presented.

A discipline that offers research possibilities is attractive to undergraduates - even though relatively few will go on to become research students. Most research leads to publications in journals or conferences (and hopefully to other outcomes) which are normally not seen by students. A notice board has been established outside the School office which features prominently the research publications of the School (in abstract form). This noticeboard is seen by all students and acts as a barometer for research activity in the departments. Interestingly, publications from the power department dominate this board.

Electrical power engineering does not generally receive much favourable publicity. However the hazards and dangers associated with the production of electrical power are widely publicised. One of the research areas within the School is renewable energy sources, and there is quite an impressive test facility associated with this research group. We have used this facility to conduct demonstrations and give talks to groups of children from local high schools. Tangible benefits from this activity are of course yet to be realised but at least the overall image of power is raised, as a discipline that is concerned about environmental problems - and has some solutions to offer.

2.3 REVISED UNITS

There has been some exciting progress over the last few years in the field of electrical power, developments such as magnetic materials, power electronics and microprocessors spring to mind. To incorporate some of these developments into final year units we simply revised syllabi and where appropriate replaced old units with new ones.[6] Examples of new material to be included:

- design of magnetic components using finite element software.
- power electronic applications
- embedded controllers (based on the 68HC11)
- power system simulation (via a lab unit)

Most of the staff in our department have maintained close contact with industry either through consulting work or professional groups such as IEAust. This allows us to set assignments which have a close relation to current industrial problems - suitably diluted for consumption by students. Such assignments are an excellent way of introducing students to real world engineering problems.

3. FUTURE DEVELOPMENTS

The trend in higher education in Australia is toward less support from central government and for institutions to supplement their income from other sources. One source of funding which is being pursued by all Australian universities is to attract overseas fee paying students. The School of Electrical Engineering at Curtin has had considerable success in attracting such students and they form a significant percentage of the student population. (20%). Given that most of these students come with some initial qualification they proceed into the course with advanced standing. Such students have also contributed to the current increase in student numbers in power. Perth is very well situated in this regard, it has the infrastructure and technology of a developed country yet has close relations with the developing countries on our doorstep. One can expect more collaboration and more students from neighbouring countries in SE Asia in the future.

4. CONCLUSION

The power course at Curtin is presently in a healthy state as a result of the initiatives described, however there is no cause for complacency. We must continue to promote the course and the activities of our department. This implies that the department should be modern, progressive and relevant to students. Basking in past glory is a sure recipe for future decline.

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TEACHING POWER SYSTEM PROTECTION AT THE UNIVERSITY OF CAPE TOWN

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The paper deals with the Power System Protection Module as a part of the Power System Course at the University of Cape Town. The student curricula and the project that has been introduced are described. Some conclusions and the student reactions are commented.

Keywords : power system, power system protection education, power system protection project

1. INTRODUCTION

In 1993 a new EEE437F course structure was introduced. In this respect the number of EEE437F lectures was reduced to 72 lectures and 2 extra credits were allocated to the project. Two different projects were introduced; one is refereed as Security and Economic Analysis of a 16' Bus Network and the second one is refereed as Power System Protection.

The Power System Protection Project comes to provide the student with the ability to understand more about how power system protection works for a given network, helps the student to develop the initiative in relay setting and co-ordination and at the same time ensures the students the protection skills.

The student mark consists of : written exam (2 papers) which counts 55 %, a project which counts 30 %, home work which counts 5 % and the oral examination which counts 10 % towards the final course mark.

The power system course became more practical and more effective.

2. POWER SYSTEM PROTECTION MODULE

The Power System Protection Module, part of Power System Course (26 lectures) covers the following topics :

- Short circuit current calculation by using symmetrical components and p.u method;

- Power System protection principles : overcurrent protection, directional, differential, distance, thermal and Bucholtz protection;

- Methods of discrimination : discrimination by time, discrimination by current magnitude, discrimination by time and direction, discrimination by both time and current magnitude;

- Instrument transformers;

- Setting the overcurrent and directional protection;

- IDMTL relay setting algorithm

- Electromechanical relays, solid state relays and microprocessor relays;

- Typical protection design: electrical motors protection, electrical line protection, power transformers protection, bus bar protection.

3. POWER SYSTEM PROTECTION PROJECT

3.1 PROJECT DESCRIPTION:

Each student has to complete the following semester project :

PROJECT DATA: A 4 bus power network is given in Figure 1. The tables 1 to 5 give the power network data. In a simplified form the power network appears as shown in Figure 2.

In order to ensure different input data for each student the total bus power at buses #3 and #4 as well as the electrical line length L1 and L2 are variable.

3.2 PROJECT STATEMENT

The study power network is supplied from two sides and is protected by distance as well as inverse time overcurrent directional relay.

The objectives of the project are :

- a. Calculate and tabulate the load flow, for :

- a1. normal regime, (normal network configuration) - base case;

- a2. line L₁ is off - CB#3 is open;

- a3. line L₂ is off - CB#7 is open;

- b. Calculate and tabulate the short circuit currents at the buses B2, B3, B4 for:

- b1. three phase short circuit :

- normal regime, (normal network configuration) - base case;

- line L₁ is off - CB#3 is open;

- line L₂ is off - CB#7 is open;

- b2. line to ground short circuit:

- normal regime, (normal network configuration) - base case;

- line L₁ is off - CB#3 is open;

- line L₂ is off - CB#7 is open;

- c. Select the CTs at the 6 line circuit breakers;

- d. Provide co-ordinated relay settings for directional overcurrent relays for the six - line circuit breakers 3, 5, 2, 6, 1, 7 associated with the power network;

- e. The settings for relay of CB 4 is known; it trips at 4 s for phase relay protection and at 2.5 s for ground relay protection; the settings of other relays are to be selected such that 1, 2, and 3 co-ordinate with 4, and 5, 6 and 7 co-ordinate with 4;
- f. Relay 4 is undirectional, and its main purpose is to backup relays 3 and 7 respectively;
- g. Student may use any of the relays whose response curves are available;
- h. The time margin is 0.3 s between relay response times of primary and backup relays at maximum fault - current flow for proper co-ordination;
- i. The student should provide the co-ordinated protective curves of the given power network;
- j. Available CTs ratios for the relays are : 50/5, 100/5, 150/5, 200/5, 300/5, 400/5, 500/5, 600/5, 700/5, 800/5, 1000/5 A;
- k. It is assumed that the phase and ground relays are connected to the same CTs.

3.3. PROJECT ASSESSMENT

1. With the Glover load flow packages find the operating point of the network by executing the load flow program (for : base case; L₁ off; L₂ off); comments on the voltage profile, network losses and current/power flow are required;
2. With the Gungor / Glover software, compute the three - phase and line to line faults at buses B2, B3, B4. Since the relays 1, 2, 3, 5, 6 and 7 are directional, it is important to know the amount of current coming from each side for faults on each bus. For this calculate and tabulate the short circuit currents with feed from LS and RS (see Fig. 2).

3. Tabulate the fault currents which are coming from each side LS (left side, Figure 2, CB #7 is open) and RS (right side, Figure 2, CB #3 is open) for faults on each bus B2, B3, B4 :
4. Select the CTs for each CB location;
5. Use the graphical time - current characteristic to find the relay settings;
6. Discuss and comment how each relay works and trips at every fault point.

4. STUDENT REACTION

The student reaction was good. The project topics were given to the students in the first week. The students became very involved in the project. Many useful direct Professor - student discussions were opened, resulting in good student training in the protection field. The student class became more active and much more involved in the course. The course achieved its objective : creating power protection skill and practical ability to the student.

5. CONCLUSION

The power system project as a part of the Power Systems Course created a good environment for teaching power system protection. At the same time the course became more attractive and efficient. The students love this course. They have the opportunity to check their knowledge practically and inventive engineering thinking.

INPUT DATA : 4 BUS POWER NETWORK

SYSTEM POWER BASE : 10 MVA

Table 1 . GENERATOR DATA

No	Type	Rating MVA	Connec.	Real Power min. MW	Real Power max. MW	React Power min. MVAR	React Power max. MVAR	V kV
G1	fossil	100	star -grnd	20	100	100	-100	18

Table 2 . GENERATOR IMPEDANCE DATA

No.	R ₁ ohm	X ₁ ohm	R ₂ ohm	X ₂ ohm	X ₀ ohm	X' _d ohm	X'' _d ohm
G1	0.001	2.04	0	0.14	0.060	0.256	0.18

Table 3 . TRANSFORMER DATA

S _r MVA	Type	Conn.	Conn.	Prim.. Vol. kV	Sec Vol. kV	R %	X %
100	tap. sec.	star- grnd.	star- grnd.	18	110	0.1	6.5

Table 4. LOAD DATA - BASE LOAD

Busbar No.	Active Power P _L , MW	Reactive Power Q _L , MVAR
B3	***	***
B4	***	***

Table 5. LINE DATA

Line	From	To	Length, Km	R ₁₀ ohm/Km	X ₁₀ ohm/Km	R ₀₀ ohm/Km	X ₀₀ ohm/Km	B ₁₀ uS/Km
L1	B2	B3	**	0.09	0.45	0.38	1.29	2
L2	B2	B4	**	0.09	0.45	0.38	1.29	2
L3	B3	B4	**	0.09	0.45	0.38	1.29	2

OUTPUT RESULTS :

Table 1
POWER FLOW LINE OUTPUT

LINE NO.	From bus	To bus	P, in kW	Q, in KVar	S, in KVA
L1	B2	B3			
L2	B2	B4			
L3	B3	B4			

Table 2
LINE FLOW VALUES

LINE NO.	From bus	To bus	Current, in A
L1	B2	B3	
L2	B2	B4	
L3	B3	B4	

Table 3
BUS OUTPUT VOLTAGES

Bus No.	VOLTAGE MAGNITUDE. p.u.	PHASE ANGLE degrees	VOLTAGE MAGNITUDE Volts	PHASE ANGLE degrees
B1				
B2				
B3				
B4				

Table 4
Fault currents, in Amps, (see Fig. 2) to the bus when a specific type of fault appears on the bus

Fault type	Bus B2	Bus B3	Bus B4
Three - phase			
Single - phase to ground			

Table 5
Fault currents in Amps, flowing from LS (L2 is off, CB#7 is open, see Fig. 2) to the bus when a specific type of fault appears on the bus

Fault type	Bus B2	Bus B3	Bus B4
Three - phase			
Single - phase to ground			

Table 6
Fault currents in Amps, flowing from RS (L1 is off, CB#3 is open, see Fig. 2) to the bus when a specific type of fault appears on the bus

Fault type	Bus B2	Bus B3	Bus B4
Three - phase			
Single - phase to ground			

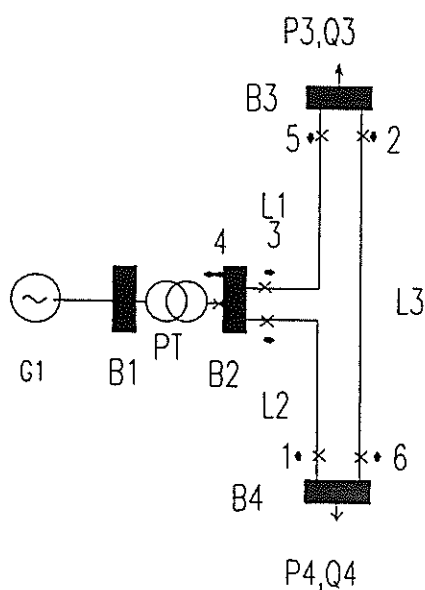


Figure 1

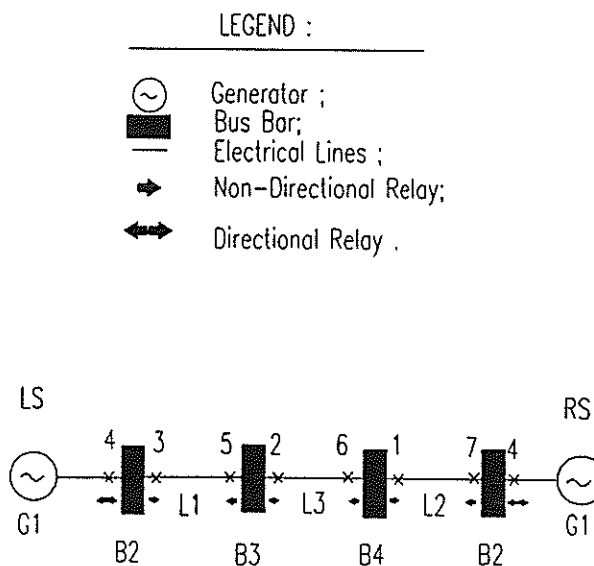


Figure 2

TEACHING POWER SYSTEMS THROUGH PROJECTS WORK

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ABSTRACT

The paper deals with the Power Systems Engineering Education at the University of Cape Town. The course physics and the projects that have been introduced are described. The student reaction, comments and some experience that has been gained after 2 years are presented.

1. INTRODUCTION

The Core Courses in Power Systems field are: Introduction to Power Systems in 3rd year (4 credits) and Power Systems in 4th year. (8 credits)

2. POWER SYSTEMS COURSE TOPICS

The following modules are covered :

- Power system components;
- High voltage;
- Electrical energy metering;
- Power system distribution;
- Power system analysis;
- Power system protection.

3. POWER SYSTEMS PROJECT TOPICS

3.1 EEE 347S PROJECT

PROJECT : POWER SYSTEM ANALYSIS AND ELECTRICAL ENERGY DISTRIBUTION

GENERAL INFORMATION

The project report counts 30% and the oral examination counts 10% of the final course mark. The project consists of three parts, namely load flow, fault analysis and electrical energy distribution.

The academic software is available through Electrical Engineering Departments' Computer network.

The power application software consists of the following modules :

- 1) Load flow & CA;
- 2) Fault Analysis (Symmetrical);

- 4) Fault Analysis (unsymmetrical);
- 5) Transient Stability Analysis;
- 6) Economic Dispatch.

The main section of the project includes the theory of load flow, fault analysis and electrical energy distribution design. The theory is supported by the output results from the programs.

PROJECT CONTENT :

Load flow

An example of a study network is illustrated in figure 1. Given the data for the base case, the students have to find the operating point of the network by executing the load flow program. One assumes initially that all tap-changing transformers have fixed taps. The students have to analyse the number of iterations required using Newton Raphson or other methods and to comment about the size of the Jacobean.. Comment on the voltage profile, network losses and the power flows are to be made.

Students have to introduce a line of equal impedance between bus B6 and bus B7. The load flow program is run again. Comments on the results are required.

Reactive compensation

The data for the heavy loading case are introduced. The students have to find the operating point for this case. One keeps the taps fixed. The students are using the load flow program again this time allowing the taps of the transformers T2 and T4 to change. They have to comment on and compare the new output data against the base case.

Comments on the results and changes are required.

The students introduce reactive compensation by installing a capacitor in the network. They have to run the program in order to analyse the effect of the reactive compensation and comment on the results.

CONTINGENCY ASSESSMENT

The students have to remove line L1 and L12 from the base case data and run the program. Comments on the output are required.

The lines previously removed are reintroduced and transformer no. T3 is removed; the students run the program and comment on the results.

Transformer no. T3 is reintroduced and transformer no. T2 is removed; the students run the program and comment on the results.

Proposed network growth

It is expected that there will be a load growth in the regions of bus B4 and B16 of respectively 120 MVA at power factor 0.90 and 40 MVA at power factor 0.95. The students enter these data and execute the program. Comments on the results are required.

Fault analysis

With the academic software, the students enter the network data and obtain the bus impedance matrix, ZBUS.

The students have to compute the three phase faults at buses B2, B11 and B16; in addition they have to compute the voltages, due to the fault, up to two nodes away from the faulted bus.

From the base case input data, the students select the relevant data for the symmetrical short circuit program and enter it into the program.

The three phase fault is selected; the students tabulate the results for the base case network. and comment on the output results.

The star grounded connection of generator G2 is changed to a delta connection. The students run the three phase fault program again and compare the results.

A. LOAD FLOW

The project should contain analysis and comments on:

- a) The load flow in power system analysis;
- b) The use of the load flow tool;
- c) The numerical algorithms used;
- d) The effect on the voltage profile, power flows and losses of the following factors :
 - i) introducing parallel lines;
 - ii) heavy loading conditions;
 - iii) reactive compensation;

- iv) tap - changing transformers;
- v) contingencies.

B. FAULT ANALYSIS

The project should contain analysis and comments on:

- a) Fault analysis.
- b) Admittance and impedance matrix.
- d) Severity of symmetrical faults.
- e) The benefit of a short circuit program for network protection analysis.

Energy Distribution Project

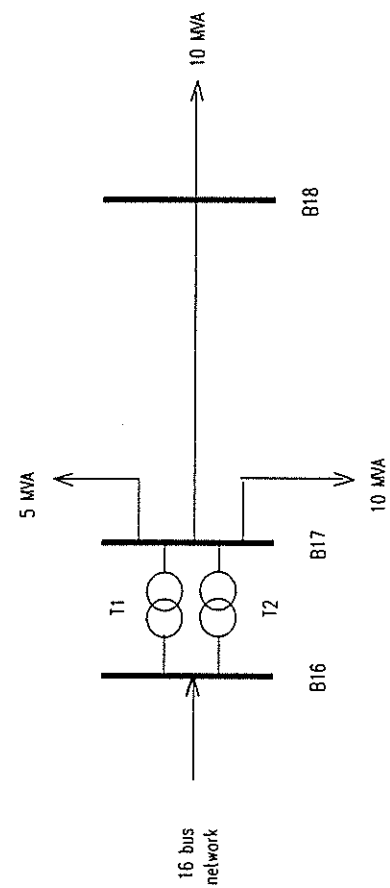
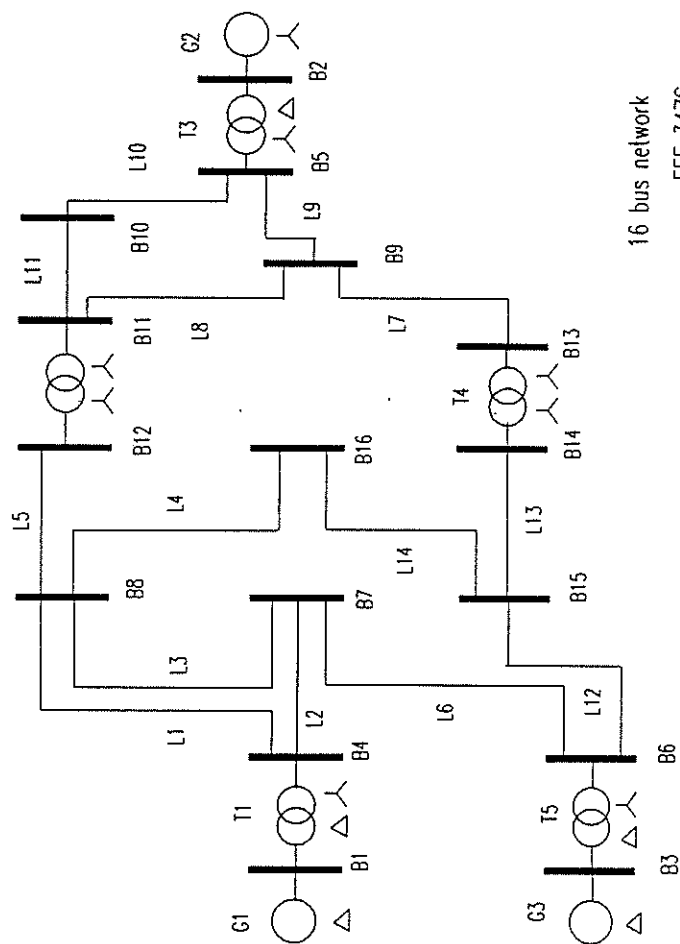
Project description :

- 1) In the base case information, the demand power at bus B17 is 25MVA at a voltage rating of 132KV. The details of this lumped load is illustrated in figure 2.
- 2) The 10 MVA energy to bus B18 is fed via an overhead line of x^1 Km long at a voltage rating of 132KV.
- 3) Between bus B16 and B17 there are two parallel connected transformers 230/132 kv.;

Project statement :

- 1) Select the power transformer rating at the power substation.
- 2) Select the line conductor cross section area;
- 3) Calculate the power transformer parameters and draw the equivalent electrical one line diagrams.
- 4) Calculate and tabulate the load flow;
- 5) Calculate and tabulate the three phase short circuit currents at the busses B16, B17 and B18;
- 6) Select the CTs, CBs and isolators;
- 7) Select and draw the power substation layout given that there should be three outputs of 2 x 10 MVA and 1 x 5 MVA.;
- 8) Draw the structure of a line feeder and power transformer including all the primary equipment.

¹ each student has a different line length.



TRANSFER FROM TECHNIKON TO UNIVERSITY

A PILOT SCHEME

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Abstract: Students transferring from a Technikon to a University and the possible acceptance of Technikon courses for credit towards a degree has been an issue for some years. This paper describes a Pilot Scheme for 1994 where students from the Cape Technikon and the Peninsula Technikon are given one year of credit and exemption towards the four year BSc(Eng) in Electrical Engineering at UCT.

1. ENTRY INTO ENGINEERING DEGREE PROGRAMMES

The majority of students who enter Engineering Degree Programmes in South Africa have completed matric. In order to qualify for entry into these programmes students are required to obtain good grades in higher grade Mathematics and Physical Science, matriculation exemption and a good overall aggregate. Most of these students are white males from a privileged educational background.

In order to increase the size of the Engineering Industry in South Africa more Engineering graduates and diplomats are required. The growth in entry to the Technikons is increasing but Universities are experiencing difficulty in attracting sufficient numbers of quality students. Industrial bursaries for first year students are declining and the numbers of white students appears to have reached saturation. The number of black students from the DET system registered for higher grade Mathematics and Physical Science remains less than 1000 in the whole country.

University bridging programmes, community colleges and other initiatives aimed at assisting educationally disadvantaged groups are starting to make a difference but at a high financial cost. It is essential to improve the mobility between Universities and Technikons. In particular, students who have excelled at the Technikons should be allowed to transfer to University and obtain credit for a block of Technikon courses and exemption from a corresponding block of University courses.

2. BACKGROUND TO TECHNIKON DIPLOMAS

The National Diploma introduced by Technikons in 1978 required a rather rigid structure with three semesters of academic work interspersed with three semesters of experiential training. This Diploma was an entrance requirement for the National Higher Diploma which required a further six months Technikon study and six months experiential training. Thus a student obtaining the National Higher Diploma required a minimum of four years -two of which were formal study.

A new National Diploma was introduced in 1992. Fig. 1 gives a comparison of the two course structure:

							ND Elec Eng	NHD Elec Eng
Old	T1	Experi- ential Training	T2	Experi- ential Training	T3	Experi- ential Training	T4	Experi- ential Training
Duration	Year 1		Year 2		Year 3		Year 4	
New	S1	S2	Experiential Training		S3	S4	New 4th year	
							ND Elec Eng	B. Tech

FIG.1

The "sandwiching" of the experiential training between periods of academic study at a Technikon represents the ideal. In practice, depending on circumstances, the experiential training may occur at different stages in the curriculum from those shown. The structural differences between the old and new courses at National Diploma level lies in the time distribution between Technikon study and experiential training. Both are 3-year diplomas, but the theory content of the new diploma is spread over two years, and the experiential training component is correspondingly reduced to one year.

The structure of the new curriculum allows sufficient flexibility in that it is possible for a student to complete the academic component in two years, followed by the one year experiential training.

Although this system is not the ideal and negates, to some extent, the co-operative style of Technikon education, it has been necessitated through lack of employment opportunities in the present economic climate.

3. PILOT PROGRAMME

Students who have completed two full academic years of study at a Technikon have been exposed to a good deal of engineering but there is a lack of basic Science courses offered in the Technikon curricula for acceptance into senior years of a University degree.

The programme to be offered from February 1994, will accept students into UCT who have completed two years of academic study from the Cape and Peninsula Technikons in Electrical Engineering. These students will be given block of Engineering credits which are equivalent to the first year of study (38 credits) and will be required to complete a block of credits (36 credits) equivalent to the second year of study.

They will be given exemption from all of the Engineering courses from the first two years (apart from Engineering Drawing) and will be required to take the required Science courses. Successful students will then continue onto the standard third and fourth years of study. It is, therefore, possible for a transferring student to graduate in 5 years after matric compared to the minimum time of 4 years (average time is 5,3 years) which students take at UCT through the regular programme after matric.

A structure for the University and Technikon programmes are shown in Figure 2 below:

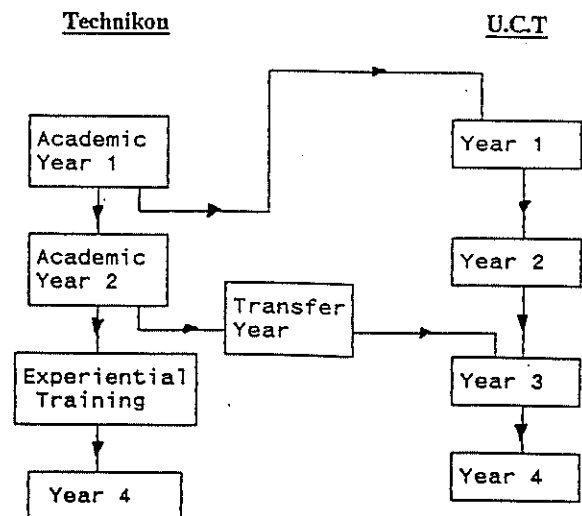


FIG. 2

4. PROPOSED CURRICULUM

The normal UCT curriculum consists of core and optional courses. The courses are given a credit rating and a student is required to acquire 144 credits plus the completion of all the core courses. The credits gained by matriculants doing the normal four year programme consists typically of 38 credits in first year, 36 in second, 35 in third and 35 in fourth.

Technikon transferees will be given a block of 38 credits equivalent to the first year and exemption of all Electrical Engineering courses in the first and second year. The subjects in their 36 credit first year are listed below.

SUBJECT	CREDITS	YEAR
Physics	9,0	First
Modelling & App. Computing	4,5	First
Engineering Drawing	4,0	First
Calculus & Vector Analysis	2,5	Second
Matrix Methods for Engineers	2,5	Second
Ordinary Diff. Equations	2,0	Second
Fourier Analysis	4,0	Second
Electrical Materials	3,0	Second
Computing for Eng. & Scientists	<u>4,5</u>	First
	36	

5. ENTRANCE REQUIREMENTS

Completion of two full years of academic study towards the National Diploma (new curriculum) and an average of at least 70% for all S3 & S4 subjects and a minimum of 75% for Mathematics II & III.

The programme is limited to students who are transferring from the Cape and Peninsula Technikons for 1994 and the number of students will be limited to not more than 10. If it proves to be successful transferring students will be accepted from any South African Technikon offering Electrical Engineering.

The percentages achieved as specified above for entry will be revised according to the success of the programme. Consideration is being given by other Engineering Departments at UCT for similar programmes in other disciplines.

GAPLESS SURGE ARRESTERS IN AC POWER NETWORKS: MONITORING THE CHARACTERISTICS OF THE NONLINEAR RESISTORS

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ABSTRACT

The basis of a technique for determining the conductance coefficient and the non-linearity coefficient of the varistors in high-voltage surge arresters is investigated. A simple and robust filtering system is used to extract the third harmonic component from the arrester's current; the data obtained agree well with those provided using a much more sophisticated and delicate instrument. Some key aspects that still need to be studied before an on-line monitor can be produced have been identified.

1. INTRODUCTION

The protection against lightning and switching surges of strategic and expensive components of a power system, e.g. transformers is of major concern to electricity supply authorities; Durban Electricity and Eskom are particularly interested in this topic. Surge arrestors, which comprise non-linear resistors mounted within a weatherhousing, are widely used for this function.

The long-term performance of these surge arrestors is questionable and so there is a growing need to investigate both ageing mechanisms and methods for monitoring their suitability for continued use.

The theoretical analysis of a proposal for making such a measurement - as based on a harmonic component of the current flowing through the arrester during normal service energization - has already been provided [1]. The current paper describes the corresponding hardware that has been constructed for such a function. The findings from making some preliminary measurements in the laboratory are presented and discussed.

2. VOLTAGE-CURRENT RELATIONSHIP

To a first approximation, the equivalent circuits of a ZnO gapless surge arrester comprises a fixed-value capacitor in parallel with a non-linear resistor. The resistive component of the current, I_r , is related to the applied voltage by the relationship:

$$I_r = kV^\alpha \quad (1)$$

where k and α are constants when I_r exceeds a relatively low value. At very low currents, the varistor acts as a linear resistor (i.e. $\alpha=1$) and so there is a transition region where α increases with rising voltage to attain its high constant value (eg $\alpha=60$). Because of this non-linear relationship, harmonic currents are produced even when the applied voltage is a pure sine wave. When the voltage waveform is distorted, complications arise - but, fortunately, these are not too serious [1] for the proposed monitor investigated herein.

Gapless surge arresters are designed so as to consume little power at the normal operating voltage. As a consequence, the capacitive current is much larger than the resistive one for the conditions under which on-line measurements can be taken. Therefore a filter and an amplifier is required.

2.1 Harmonic Component of I_r

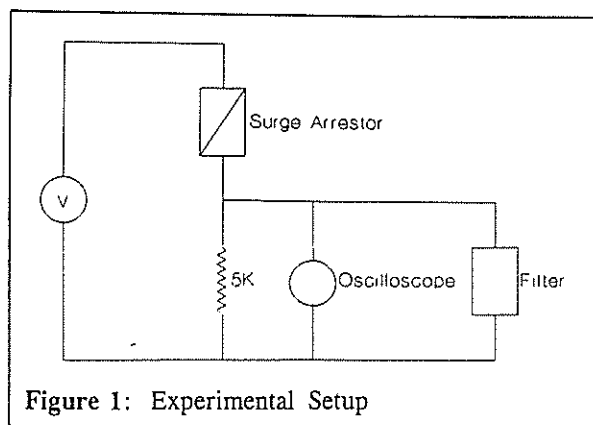
Using MATLAB [2] it was found that when plotting the third harmonic of the resistive current in decibels, it varied linearly with the log of the input voltage V . It was further found that α is directly proportional to the slope, m of this linear relationship; i.e

$$\alpha = \frac{m}{20} \quad (2)$$

The factor 20 in equation (2) originates from the fact that the third harmonic is recorded in decibels. It was also found that only one specific value of K would give the an exact line passing through two points.

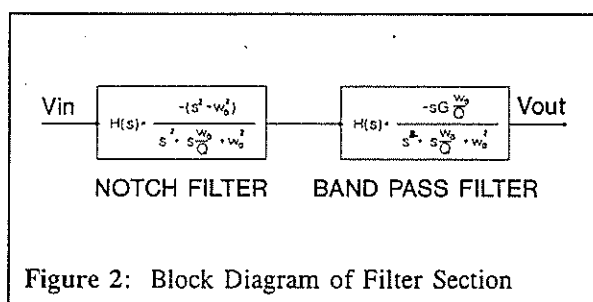
3. EXPERIMENTAL SETUP

An electrical filter was designed to capture the third harmonic of the surge arrester current. Figure 1 shows the experimental setup. The digital oscilloscope used (Hewlett Packard 54510 A) has an FFT function, which enabled the harmonic content of the current to be recorded directly. These values could then be compared with those given by the filter circuit thereby investigating its effectiveness.



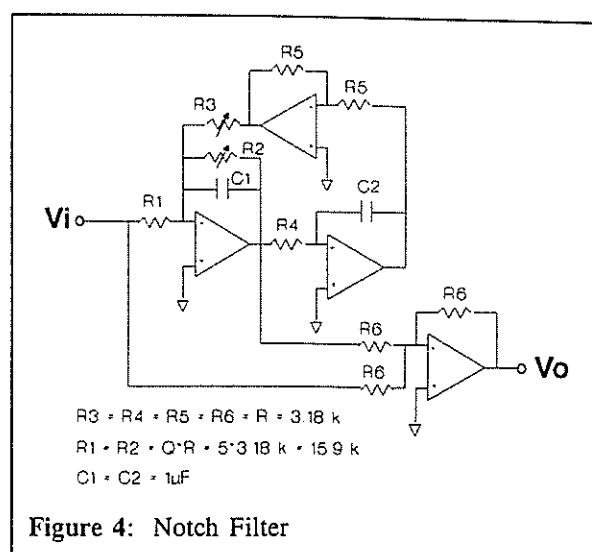
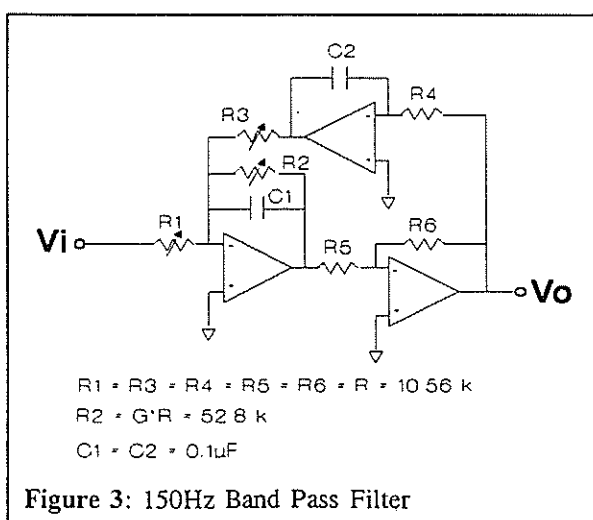
4. FILTER CIRCUIT

Since the capacitive component of the total current is far greater than the resistive current, the fundamental of the resistive current is much larger than the magnitude of the third harmonic. It was therefore decided to first remove the fundamental from the resistive current. This was achieved using a 50 Hz notch filter. The output of this filter was then fed into a band pass filter centred around 150 Hz. A block diagram showing the total filter is shown in figure (2)



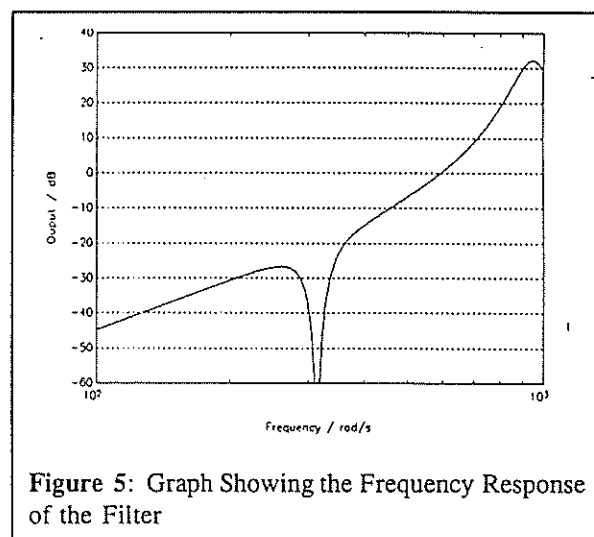
The common biquad filter [3] was chosen for both the notch and band pass filters for the following reasons;

- * it can be used to realise both the band pass and the notch characteristic,
- * the Q-factor can be easily tuned,
- * the gain can be easily adjusted,



- * the centre frequency can be easily tuned without any effect on the Q-factor
- * has low sensitivity to component variation.

The band pass filter is shown in figure (3) and the notch filter is shown in figure (4). Figure 5 shows the frequency response of the total filter.



5. TEST RESULTS

The experimental setup shown in Figure 2 was used on four different surge arrestors, each energized over a voltage range from 10 to 15 kV, to determine the third harmonic of the resistive current. Once the third harmonic had been recorded the corresponding values of α and K were calculated as the voltage increased using the method described earlier. Tables (1)-(4) show the results of the tests performed and the corresponding values from the calculations. In these Tables, α_1 and K_1 were calculated using the filter third harmonic values and α_2 and K_2 were calculated using the oscilloscope third harmonic values. Note: each set of α and K values were calculated from results from two voltage levels that were only 0.5kV apart.

V _{in} /kV	Filter /dB	Oscil scope /dB	α_1	K ₁	α_2	K ₂
10	-20.31	-20.94				
10.5	-18.75	-18.44	1.0	4.1e-9	2.1	2.3e-15
11	-18.44	-16.88				
11.5	-16.88	-16.255				
12	-14.69	-14.69	2.2	1.7e-15	1.3	2.5e-11
12.5	-13.13	-12.50				
13	-11.25	-9.69				
13.5	-8.44	-8.13	3.4	1.1e-20	3.2	9.3e-20
14	-6.25	-5.00				
14.5	-2.81	-2.50				
15	0.63	0.63	5.04	2.1e-27	4.1	1.3e-23

Table 1: Surge Arrestor A

V _{in} /kV	Filter /dB	Oscil scope /dB	α_1	K ₁	α_2	K ₂
10	-23.44	-24.06				
10.5	-22.75	-24.06	1.5	1.6e-12	2.1	2.0e-15
11	-20.63	-20.00				
11.5	-18.75	-19.06				
12	-16.81	-17.50	2.2	1.1e-15	1.4	2.5e-12
12.5	-15.63	-14.06				
13	-13.75	-13.44				
13.5	-10.63	-9.69	3.6	1.3e-21	4.0	2.3e-23
14	-8.44	-7.50				
14.5	-5.63	-6.25				
15	-2.19	-1.56	4.5	1.4e-25	4.3	1.3e-24

Table 3: Surge Arrestor C

V _{in} /kV	Filter /dB	Oscil scope /dB	α_1	K ₁	α_2	K ₂
10	-21.57	-21.25				
10.5	-20.31	-18.75	1.5	1.8e-12	2.0	1.1e-14
11	-18.75	-17.50				
11.5	-16.11	-15.94				
12	-15.63	-14.69	1.8	6.0e-14	1.6	4.1e-13
12.5	-13.13	-12.81				
13	-10.94	-10.94				
13.5	-8.44	-7.50	3.4	1.1e-20	4.0	2.6e-23
14	-5.94	-5.00				
14.5	-2.81	-2.50				
15	0.94	0.94	5.0	2.1e-27	4.3	1.5e-24

Table 2: Surge Arrestor B

V _{in} /kV	Filter /dB	Oscil scope /dB	α_1	K ₁	α_2	K ₂
10	-24.06	-22.81				
10.5	-20.94	-20.63	2.3	4.0e-16	2.3	4.2e-16
11	-19.69	-18.44				
11.5	-17.50	-17.81				
12	-16.56	-15.31	1.8	5.7e-14	1.8	6.1e-14
12.5	-15.00	-12.81				
13	-11.88	-10.63				
13.5	-9.69	-8.44	3.4	1.1e-20	3.0	6.9e-19
14	-6.88	-6.25				
14.5	-4.06	-3.44				
15	-0.31	0.63	4.8	1.8e-26	5.0	2.1e-27

Table 4: Surge Arrestor D

6. DISCUSSION

The computed values for α provided in Tables 1 to 4 are shown together in Figure 6 for the four arresters. It is very encouraging to find that the results obtained using a simple notch filter plus band pass filter are in close agreement with those provided by the FFT function of a high-quality oscilloscope.

Although such values are not all that steady, they do follow the expected trend of increasing as the test voltage increased. The reason for the lack of smoothness in the curves is not yet properly understood but may be due, in part, to the rapidly changing voltage levels that are known to occur frequently in the laboratory power supply. Further, the test circuit was constructed on a breadboard trainer using cheap components. It is expected that once the

circuit has been reconstructed using high-quality components, the results will be far more accurate.

Being able to determine α and K from voltage levels that are fairly close together is of practical importance. That is, the voltage regulation that occurs on distribution and transmission circuits can be employed to good effect when making on-line measurements from the arresters current.

The main aspects that next need to be investigated using this monitoring circuit are the repeatability of the findings for the same arrester and the correlation between the α and K values determined at normal service conditions with those pertaining when large surge currents are being conducted.

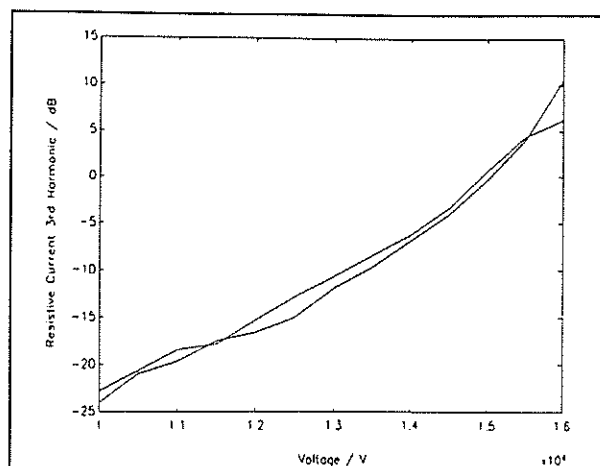


Figure 6: Graph showing how the predicted values of α and K produce a curve which fits those recorded.

7. CONCLUSION

A simple notch filter plus a band pass filter can readily extract the third harmonic component of the resistive current from the total 50 Hz current flowing through a high-voltage ZnO surge arrester, even when this component is much smaller than the capacitive component. Such values agree closely with those obtained using the FFT function of a high-quality oscilloscope.

The information contained in this harmonic current can be used to calculate the conductance coefficient and the non-linearity coefficient of the varistors.

These characteristics of a gapless surge arrester can be determined by using two voltage levels that need be no more than 500 volts apart. Surge findings are encouraging for the development of an on-line monitor because the voltage regulations of power system readily provides such changes.

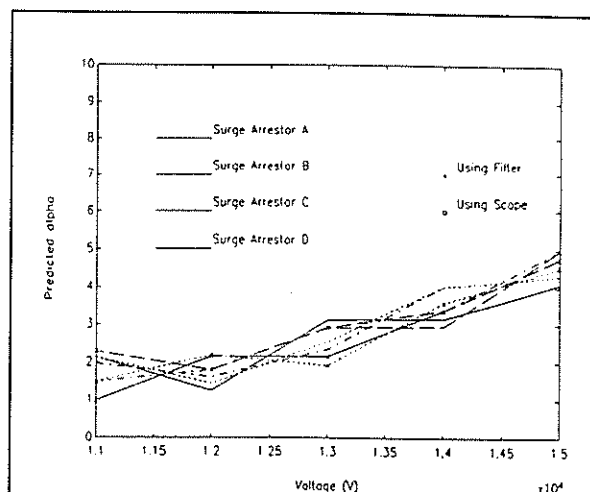


Figure 7: Graph showing the predicted values of α for the surge arresters.

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