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Potchefstroom University of Christian
Higher Education**

19-20 January 1999



**Potchefstroomse Universiteit
vir Christelike Hoër Onderwys**

S. Mouton

**SAUPEC '99
EIGHTH SOUTHERN AFRICAN UNIVERSITIES POWER
ENGINEERING CONFERENCE**

19-20 January 1999

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Tuesday 19 January 1999

(Natural Sciences Building; Level 1)

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THE USE OF SERIES-INJECTION AS AN ALTERNATIVE TO TAP CHANGERS IN LOW AND MEDIUM VOLTAGE NETWORKS

1

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Abstract - Series voltage injection is most often used for voltage compensation. In this study the use of series voltage injection is expanded to include on-load tapchanging in power systems. This expansion provides an improved voltage regulation system by integrating a discrete tapchanging transformer and series voltage compensator into a single unit. This combination allows the transient response of the tap changer to be eliminated due to the controllability of the series injected voltage. A PWM converter is used as the series voltage source supplying the required compensating voltage.

Keywords - Series voltage injection, Electronic tap changer, SWER

I. INTRODUCTION

Due to the voltage drop across the impedance present in transmission lines the voltage level supplied to the customer does not always conform to specifications [1]. In low and medium voltage networks tap changers are used to maintain correct system voltages.

With the improved power capabilities of semiconductor switches it has become a viable option to place a variable voltage source in series with the supply voltage to compensate for the voltage drop across the line impedance [3][4]. This variable voltage can be supplied by a PWM-switchmode converter and would thus effectively replace the tap changer. An advantage of this proposed topology is the controlled dynamic compensation ability in addition to what can be achieved by tapchanging. The use of series voltage injection as an alternative to tap-changers is aimed at Single Wire Earth Return (SWER) systems as a means of improving quality of supply and subsequently the number of users connected to a single feeding line [10].

II. TAPCHANGING

Conventional tapchanging transformers change the number of turns on the primary or secondary winding of the transformer to compensate for output voltage loss, across the line impedance at increased load. Only step voltage changes are possible because of the limited number of taps, which in turn is limited by the integer value of the turns on a given winding. The changing of taps on a transformer can be done while running at full load or during an off-load period. The latter generally only takes place during isolated intervals and the power is required to be interrupted every time a tap is changed. Off load voltage compensation is

therefore only used when the distribution systems' parameters change. Changes in network voltage levels normally take place during a significant increase in load or altered line impedance. Daily fluctuations in network voltage levels can not be compensated for by off load tap changers. On load tap changers can be used to compensate at more regular intervals. These tap changers however suffer from several disadvantages. The change of a tap introduces transients into the network due to the stepchange in voltage. Non-ideal contact phenomena and the repeated connecting and disconnecting of high voltages causes arcing that erodes contacts forming impurities in the transformer oil. High levels of maintenance are therefore required on tapchanging transformers [13].

The high maintenance of on load tap changers can be offset to some extent with the use of solid state tap changer implemented with thyristors. Before the tap changer can change to the next tap a sufficient voltage error level must be reached causing an incorrect voltage level to be maintained for an unknown period of time. Fast error correction could be helpful to minimise the probability of shutdown in sensitive equipment due to undervoltage conditions before tapchanging takes place. With the use of high-power semiconductor switches, series voltage injection by switchmode converters become a viable option at the power levels required for power distribution in low and medium voltage networks. This type of compensation, although more complex in nature, has significant advantages such as improved dynamic response which will be discussed. It also becomes possible to compensate for harmonic voltage distortion in addition to the tapchanging function. The faster dynamics of such a system can however lead to network stability

flux density in the air gap compared with that of the buried system. In this case the average air gap flux density is 0.49T.

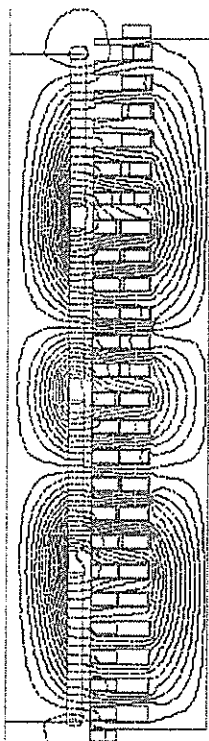


Figure 7 Flux plot of the surface mounted system in the tubular LSM

The reason that it is lower than that of the single-sided system is that there is saturation in the back iron of the secondary. This occurs where there is a concentration of flux lines and can be seen in Figure 7. In a tubular system, the area is proportional to the radius squared. So a small reduction of the diameter decreases the area considerably and the back iron becomes saturated. If the diameter is increased, the volume increases considerably, thus affecting the force to weight ratio. Thus, an optimum needs to be reached. It has not been done in this case as the secondary has been kept at the same diameter as that of the buried magnet configuration for comparison purposes.

4. CONCLUSION

It appears that the tubular LSM has an advantage of an improved force to weight ratio over both the buried and surface mounted magnet systems in the single-sided LSM. This is mainly due to the fact that the air gap area (lw) has been increased by twice that of the single-sided system, and the mass of the mover has been kept constant. Investigation into the tubular LSM has confirmed work done on comparisons between buried and surface mounted magnets [2]. It is envisaged that a system will be built and tested, and

that the results presented here from finite element analysis will correlate with those from practical testing.

5. REFERENCES

- [1] R.J Cruise and C.F. Landy, *Linear Synchronous Motor Propelled Hoists*, Symposium on Power Electronics Electric Drives and Advanced Electric Motors, Capri, June 1996
- [2] C.G. Jeans, A. Hansa, R.J. Cruise and C.F. Landy, *Comparison between surface mounted and buried permanent magnets in linear synchronous motors*, Symposium on Linear Drives for Industry Applications, Tokyo, April 1998
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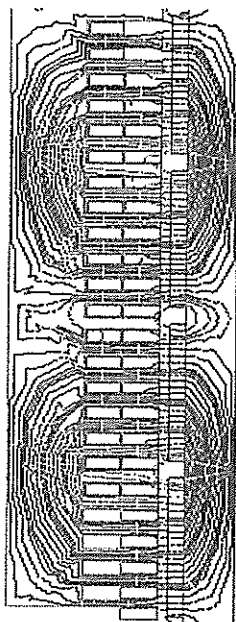


Figure 4 Flux plot of the surface mounted system in the single-sided LSM

However, the surface mounted single-sided system relies on a substantially thick back iron in the secondary design. This is mainly due to the high flux density area between adjacent magnets. This is unavoidable as all the flux from the one pole of a magnet flows into the one pole of one adjacent magnet (Figure 2). The reason why the flux is not distributed to adjacent magnets on both sides is an area for further investigation.

3.3 Buried Magnet Design in the Tubular LSM

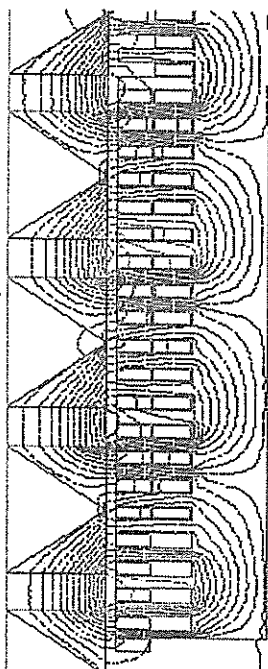


Figure 5 Flux plot of the buried system in the tubular LSM

For buried magnets in the single-sided system, the depth of the iron (pole shoe) is the same as the depth of the magnets themselves. This is the same for the tubular design, as the diameter of the pole shoe and the magnets are the same. But the flux in the buried magnet case is more evenly distributed in the pole shoes compared with that of the back iron in the surface mounted magnets. Thus it is possible to concentrate the flux in a comparatively smaller amount of steel. This can be seen in Figure 5. By reducing the amount of steel in the secondary, the overall weight of the mover is reduced and hence there will be an improved force to weight ratio.

Also, in the tubular design, the flux has nowhere else to go but through the air gap and into the primary stator. Thus there is a more efficient use of the magnet material, and very little leakage flux. However, as in the buried single-sided system, the average air gap flux density is fairly low and is approximately 0.45T. There is an improvement due to the fact that there is less leakage and more magnet material.

Another aspect in the buried magnet design, is that of magnet length optimisation. In the single-sided LSM the magnet length was 33% of the pole pitch [3]. In the tubular design, the maximum force attained for varying magnet lengths was investigated. From this investigation it was apparent that as the magnet length increased some saturation effects were introduced. This was verified in the investigation of the average flux density in the air gap with varying magnet lengths.

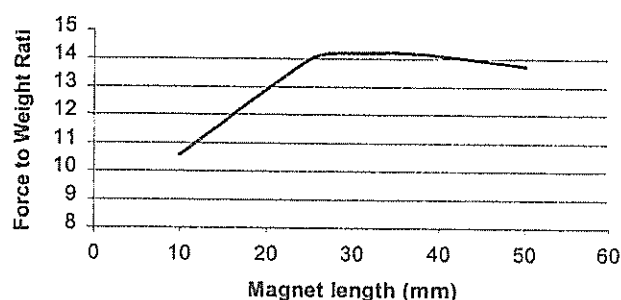


Figure 6 Force to Weight ratio versus magnet length

Owing to the geometry of the model, as the magnet length is increased, so does the effective volume and hence mass of the mover. So if a force to weight ratio versus magnet length is plotted (Figure 6), it shows the optimum magnet length. It must be noted that the pole pitch is 114mm

3.4 Surface Mounted Magnets in the Tubular LSM

For surface mounted magnets in the tubular LSM, the system has the same advantage as in the single-sided case, as this magnet configuration produces a higher

mover mass = 22kg). With the tubular LSM, there is a maximum force of 2972N and the weight of the mover is 20.8kg, which results in a weight ratio of 14:1. Lastly, the maximum force achieved by the surface mounted magnets in the tubular LSM is 3056N. However, the mass of this system is 28.1kg. Thus, the force to weight ratio is 11:1. All these force values were obtained by using finite element analysis. In the case of the single-sided buried system Nd-Fe-B30 magnets were used, and in the surface mounted and the tubular configurations Nd-Fe-B35 magnets were used. These calculations and hence the force to weight ratios were obtained under the same current conditions.

The improvement in the forces developed of the tubular design over the single-sided design for the two cases of magnet arrangements can be explained. Firstly, due to the nature of a tubular design, there is a flux concentration towards the centre of the tube. However, this deals only with flux leakage and the distribution of flux in the movers. This is of relative minor importance compared with the next reason, and that is of the difference in the air gap area. This is concerned with the maximum force generated, and can be described by the following equation:

$$F = \overline{BQ}lw \quad \dots(1)$$

Where B = Average flux density (T)
 Q = Electrical loading (A/m)
 l = length (m)
 w = width (m)

Now since both the single-side LSM and the tubular LSM have the same primary (Stator and windings) Q and l are also the same. Since

$$Q = I_{ph} \times \text{cond} / \text{slot} \times \text{slots} / l \dots(2)$$

Where I_{ph} = Current per phase (A)

For the single-side system $w = 152\text{mm}$ (two single-sided sections, each 76mm) and for the tubular system $w = 2 \times r = 314.16\text{mm}$. Thus if the average flux density in the air gap was the same for both systems the force generated by the tubular LSM should be approximately twice that of the single-side LSM.

It must be noted that, in the finite element analysis, the simulations for both the single-side LSM and the tubular LSM designs have been done in a two dimensional environment. For the single-sided LSM simulations, the models were drawn in the xy-plane, and were assumed to extend one metre into that plane. Once solved, the results were scaled accordingly. For the tubular LSM, the models were drawn in the rz-plane and this two-dimensional model is assumed to revolve around the z-axis. The z-axis in Figure 5 is the light grey line on the left-hand side of the diagram.)

3.1 Buried Magnet Design in the Single-Sided LSM

In the secondary of single-sided LSM design with the buried magnet configuration, there appears to be a large amount of leakage flux on the side opposite to that of the primary section (Figure 3). Although most of the flux does flow through the air gap and into the primary stator it still has to pass through two 5mm air gaps to complete the flux path. With each magnet being only 38mm long, there is a large amount of leakage flux especially with the two magnets on the ends of the mover. Due to these reasons the average flux density in the air gap is only 0.35T, and is the system that produces the least force.

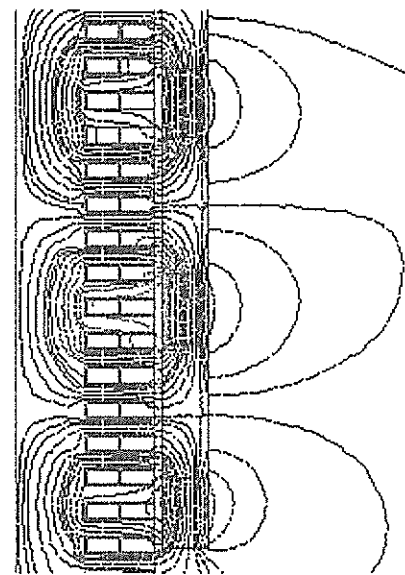


Figure 3 Flux plot of the buried system in the single-sided LSM

3.2 Surface Mounted Magnets in the Single-Sided LSM

The reason for an improvement with the surface mounted magnets in the single-sided system is that there is a reduction in leakage flux as all the flux flowing from the one pole is forced into the air gap (Figure 4). However, the flux from the opposite pole is provided with an easy path to the complementary pole of the adjacent magnet by means of a back iron. It would be beneficial to use this flux in say, a double-sided system for greater efficiency, as there would be no need to use a back iron.

In the single-sided system it is apparent that the surface mounted magnets provide an improved performance compared to that of the buried magnet system. This is due to the higher average flux density (0.55T) generated in the air gap with the surface mounted configuration.

COMPARISON BETWEEN SINGLE-SIDED AND TUBULAR LINEAR SYNCHRONOUS MOTOR TOPOLOGIES

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Abstract - A comparative study between single-sided and tubular linear synchronous machine topologies is presented. The single-sided linear synchronous machine has already been built and tested, and physical test results compare favourably with finite element analysis. On this basis, the tubular linear synchronous machine has been designed purely through finite element analysis. The two cases that are investigated are those of a buried magnet configuration, and surface mounted magnets. This comparative study concentrates on the force to weight ratio that the motors can achieve as this an important performance characteristic of linear motors.

Keywords: single-sided and tubular linear synchronous motors, buried and surface mounted magnets, force to weight ratio.

1. INTRODUCTION

In South Africa, all gold mines use rope hoists to transport men and materials to and from the site of where the gold is actually mined. Due to stringent safety in the mining industry, a safety factor of 8 is incorporated in the diameter of the hoisting rope. This means that after a certain depth the actual weight of the rope is more than a fully loaded cage [1]. With the gold bearing reef at 3000m, it requires a sub-vertical shaft to be sunk underground in order to reach the reef. With sub-vertical shafts, the travel time of both men and material increases, reducing productivity. With gold now being discovered at levels of five thousand meters, new ways of transport need to be investigated. With a linear synchronous machine (LSM) the depth is not limited. The only limiting factors deal with rock mechanics, and thermal considerations. Also using a LSM, the diameter of a shaft can be decreased, reducing the cost and time of sinking a shaft.

Thus, the Force to Weight Ratio (FWR) has become an important factor in the design of a LSM. The force to weight ratio is the quotient of the maximum force obtained and the weight of the secondary. In the case of both the single-sided LSM and the tubular LSM, the secondary is the mover.

2. BACKGROUND

The single-sided LSM has a long primary, and a short secondary. The long primary is made up of a three-phase double layer winding, with 2 slots/pole/phase. It has been short chorded by one slot to reduce the negative effects of harmonics. This primary winding has two different secondary configurations. Both consist of permanent magnets, one with buried, and the other with surface mounted magnets. The tubular LSM has been designed with the same primary winding configuration. This is purely for comparative purposes.

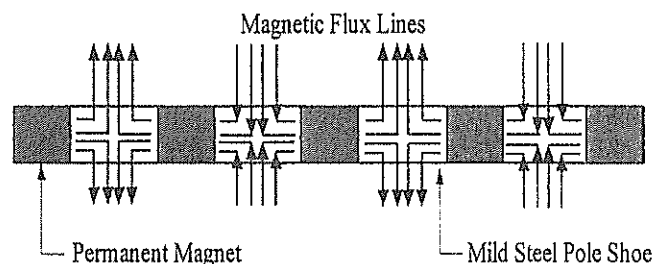


Figure.1 Buried magnet arrangement in the single-sided and tubular LSM [2]

Figure. 1 shows the four pole buried magnet configuration. As the flux lines indicate, the magnets are orientated such that the mild steel blocks act as pole shoes. The other design can be seen in Figure 2. The magnets are mounted onto a back iron and the magnets form the poles.

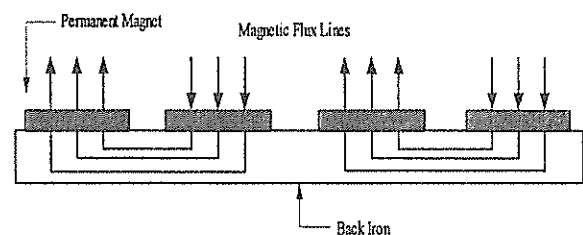


Figure 2 Surface mounted magnet arrangement in the single-sided and the tubular LSM [2]

3. COMPARISON OF FORCE TO WEIGHT RATIOS

It was found, for the single-sided LSM, with the buried magnet configuration, that a maximum force of 717N was achieved. With the mass of the mover being 19.2kg, the force to weight ratio is only 4:1. Whereas for the surface mounted magnets a force to weight ratio of 7:1 was achieved (Maximum force = 1681N;

Phase voltages of 71 V peak-to-peak and line voltages of 118 V peak to peak were obtained at 500 rpm. The mechanical system used tends to overheat and seize. This factor limits the operating time span of the generator.

The results of the system tested are encouraging. It is recommended that a tubular generator (with six stator sections arranged in a tubular formation) using ring permanent magnets be built and tested. Also a Renault two stroke engine be used to produce the linear motion is recommended.

7. REFERENCES

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4. RESULTS

The results obtained from the system are tabulated below:

The phase sequence is Red, Blue, Yellow. All voltages are given as peak-to-peak values.

Table 1 Phase voltages generated

Shaft speed (rpm)	Coil voltage (V)	Red (V)	Blue (V)	Yellow (V)
200	4	34.8	32.4	32
500	8.8	72.4	71.2	70

Table 2 Line voltages generated

Shaft speed (rpm)	Red-Blue (V)	Blue-Yellow (V)
200	53	50
500	118	116

Based on Table 2, a curve of voltages versus motor shaft speed is plotted in Figure 4.

The generator built and tested, is a three phase generator. When the mover with the permanent magnets moves in one direction, the phase sequence generated, will be the reverse of the phase sequence generated when the mover moves in the opposite direction. The voltage generated will have to be rectified to dc before it can be stored in the batteries. The reversal of phase sequence does not pose a problem for a three phase rectifier.

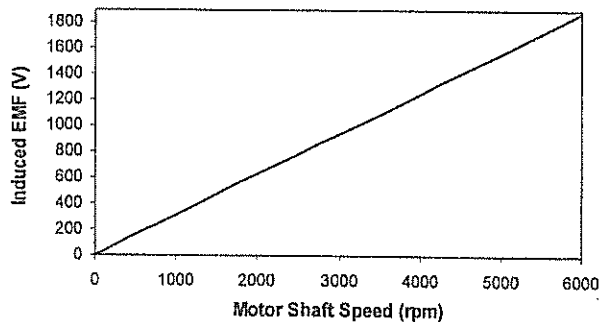


Figure 4 Curve of voltage generated versus motor shaft speed

The reversal of the phase sequence is not a problem for the HEV application for which the linear generator is intended.

5. Tubular generator design

The size of the generator, as it stands, is too big to install in a 1400 cc vehicle. It is recommended that a tubular generator (with six stator sections arranged in a tubular formation) be built and tested (Figure 5).

Two ring permanent magnets will be placed on the piston shaft. The magnets will be surrounded by the six stator sections. It is further suggested that each stator section be of a two-pole, three phase winding, 5/6 chorded design.

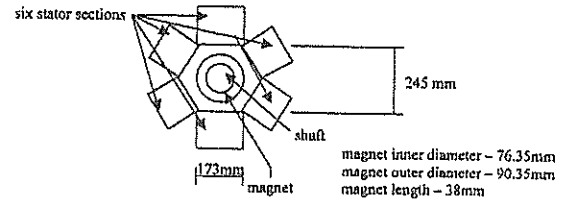


Figure 5 Drawing of tubular generator

Theoretically, the expected phase voltage (peak-to-peak) is:

58.35 V at 200 rpm
145.86 V at 500 rpm
1166.78 V at 4000 rpm
1750.23 V at 6000 rpm

The induced emf obtained from the tubular generator, is greater than the values obtained in the double sided system, at the same motor shaft speeds. This is enough to charge a 400 V battery.

It is further recommended that a Renault two stroke engine be used to simulate the linear piston motion. The linear generator is being designed to function on such a motor. This system will not have the same problems (namely overheating) as the existing mechanical system that is used to simulate the piston motion.

6. CONCLUSION

The feasibility of a linear generator for a HEV was evaluated. It is established that the linear piston motion will drive the permanent magnets in a linear motion. The generator tested is a double-sided system. Each stator section has four poles and the windings are 5/6 chorded. Four NdFeB35 permanent magnets were used.

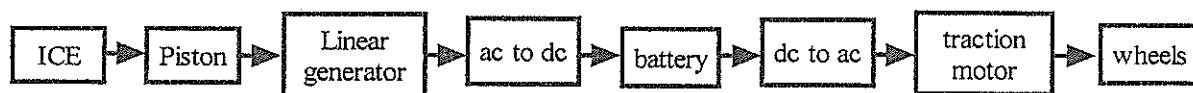


Figure 1 Hybrid electric vehicle in series connection

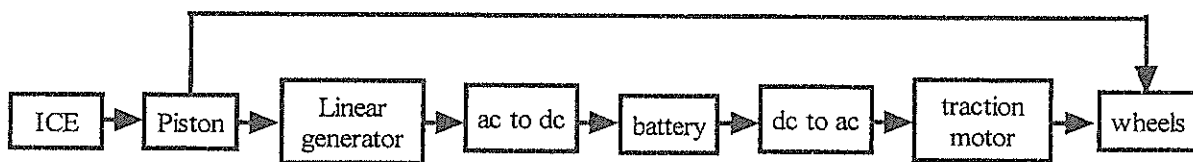


Figure 2 Hybrid electric vehicle in parallel connection

For rotary generators, the linear motion of the piston is converted to rotary motion (this motion is used to drive the rotary generators). The generators that are used in most Renault HEVs are rotary [1]. In this project the feasibility of linear generators is investigated. The advantage of the linear generator is that the linear motion of the piston can be used directly to generate charge. The conversion of piston linear motion to rotary motion is not needed, hence efficiency can be increased.

3. DESIGN

3.1 Production of linear motion

The linear motion of the piston is reproduced using a reluctance synchronous motor, a cam shaft and bushings as shown in Figure 3. The reluctance synchronous motor used, provided variable output shaft speeds. The motor has a rated output power of 5.5 kW. The motor on no-load can reach a maximum speed of 3000 rpm. The starting torque is 62.78 N.m. This is, however, less than the 6000 rpm needed to simulate a piston.

The motor shaft drives a cam shaft which produces linear motion in shaft r3 (Figure 3). The permanent magnets are connected to shaft r3. The bushings ensure that the shafts r3 and r5 as well as the permanent magnets move in a linear direction with no lateral or vertical movement. No angular rotation of r3 was observed when the system was built and tested. The gap between the magnets and the stator is maintained at 11 mm using non-magnetic spacers.

A limitation was discovered when the system was running. This limitation manifested itself as a temperature rise. The space between the bushing b1 and the shaft r3 was made 0.5 mm to minimise any lateral or vertical movement. When the shaft moved within the bushing, the temperature of the shaft and the bushing increased due to the friction between the two. The rise in temperature caused the shaft r3 to

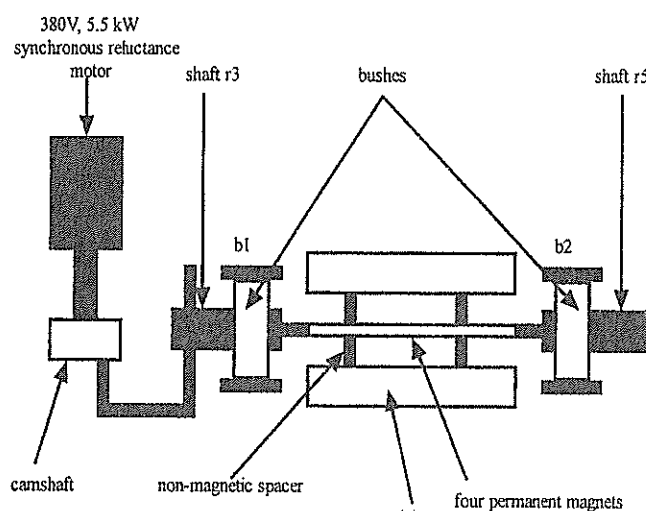


Figure 3 Top view of mechanical system that simulated the piston motion

expand and the bushing b1 to contract, thereby reducing the space between the two significantly and causing the system to seize. As shaft r3 and bushing b1 started to expand, shaft r3's movement, and hence the permanent magnet movement, was not smooth. As a result, the system could not be run at high frequencies and/or for long periods of time. The maximum speed reached was 500 rpm.

3.2 Linear generator design

The generator built, is a double-sided system using four rectangular permanent magnets. The magnets (of type NdFeB35) are attached to the shaft as shown in Figure 3. The gap between the magnets and the stator is 11 mm.

The stators are of a four pole design. The pole pitch is 114 mm. The winding is double layer and 5/6 chorded.

LINEAR GENERATOR FOR HYBRID ELECTRIC VEHICLE

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Abstract: The feasibility of using a linear generator in the place of a rotary generator for a Renault hybrid electric vehicle was investigated. The generator tested is a double-sided system. Two, four-pole, three-phase stator sections were used. Four NdFeB35 permanent magnets were attached to a shaft that simulated linear, piston motion. The phase and line voltages produced are 71 V peak-to-peak at 500 rpm, and 118 V peak-to-peak at 500 rpm respectively. The phase and line voltages obtained are encouraging. It is recommended that a tubular generator with six stator sections (each stator has two poles and three phase winding) be constructed and tested. It is further recommended that the mechanical system that is used to produce the linear piston motion, be changed. A Renault two stroke engine be used instead (the linear generator is to be designed for such an engine).

Keywords: linear generator, hybrid electric vehicle.

1. INTRODUCTION

The focus of this project is on the design, construction and performance of a linear generator with the aim of installing it in a Renault hybrid electric vehicle (HEV).

The hybrid engine under consideration is in series connection (Figure 1). This means that the internal combustion engine (ICE) drives a piston. The linear motion of the piston is converted to electrical energy by a linear generator. The electrical energy generated is stored in a battery (the battery is a 400 V 40 A battery). The battery then supplies a traction motor, which drives the wheels of the car. The electrical energy is generated in ac form. It must be converted to dc by a rectifier before it is stored in the battery. From the battery the electrical energy must be converted from dc to ac by an inverter before it can supply the traction motor.

2. BACKGROUND

HEVs are vehicles that have both an electric motor and an internal combustion engine.

HEVs have distinct advantages over vehicles powered by ICEs only. The consumption of litres of fuel per kilometer is less. An average vehicle powered by ICE does 500 km on 50 litres, HEVs can do 500 km on 17 litres [1]. The speed and distance performance of the HEV is comparable to a vehicle powered by ICE only [1]. Because less petrol is used per kilometer in a HEV, the harmful emissions from an HEV is less compared to that of a vehicle powered by ICE only.

Electric vehicles (EV) are powered by electric motors only (there is no ICE in these vehicles). The motors are run from a battery. The batteries used are bulky and take up a lot of room in the vehicle. Recharging of the batteries takes a long period of time [2]. The electric vehicle performance is not comparable to vehicles powered by ICE. A top speed of 120km/h has been recorded for EVs [2]. The battery used for HEVs need not be bulky, as the amount of charge stored is not as much as the amount stored in the batteries used for an EV. Recharging of HEV battery occurs whilst the vehicle is in motion. If at any time the battery of an HEV needs to be recharged, the driver should press the accelerator. This causes the piston to oscillate, hence driving the generator to produce charge. This charge is then stored in the battery. Therefore recharging the battery in an HEV does not take a long period of time as compared to EVs.

Series and parallel configuration HEVs can exist. In a series case (Figure 1), only the traction motor powers the wheels of the vehicle. In a parallel configuration, the piston is also connected to the wheels of the vehicle (Figure 2). These figures can be seen on the next page.

A control system is used to ensure that the combination of piston and traction motor provide adequate power to the wheels of the vehicle. When braking, or slowing down, the vehicle uses regenerative braking to recover energy. An example of a parallel configuration is the Toyota Prius [3]. During start up this vehicle runs on battery power. When the vehicle runs up to speed, both the traction motor and the piston power the vehicle.

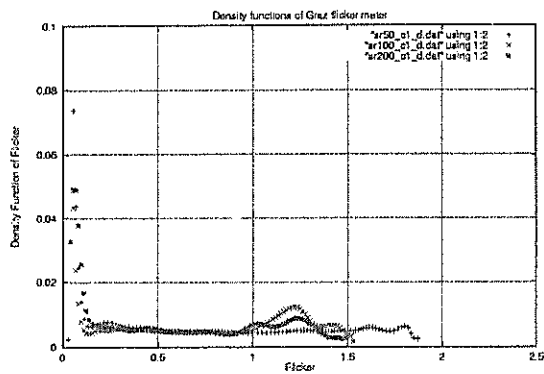
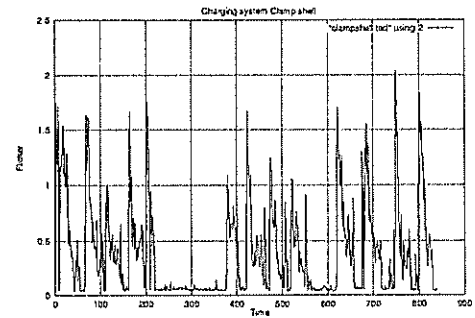
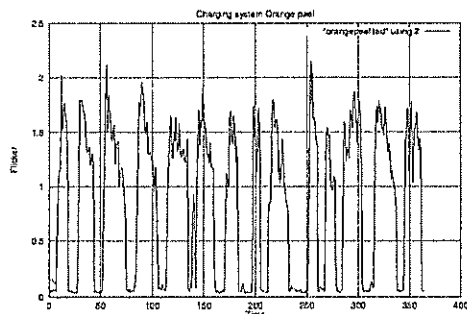


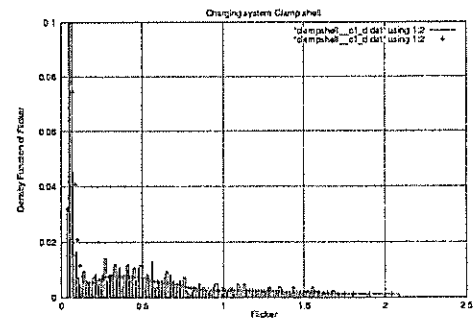
Fig. 7. Smoothed density function comparison of SR



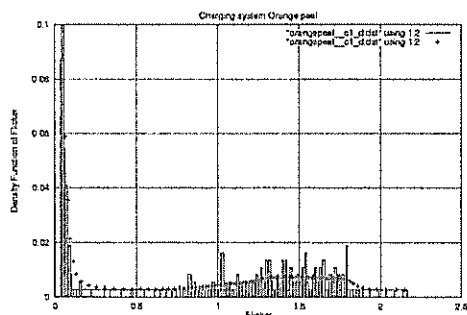
(a) Flicker vs. Time plot



(a) Flicker vs. Time plot



(b) Flicker density function



(b) Flicker density function

Fig. 8. Charging system Orange peel

and prediction. *Laborelec, Electrabel & UCL, Belgium, 1993.*

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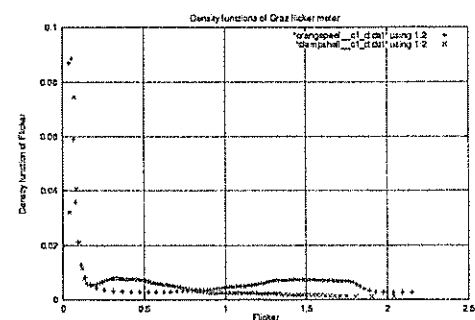
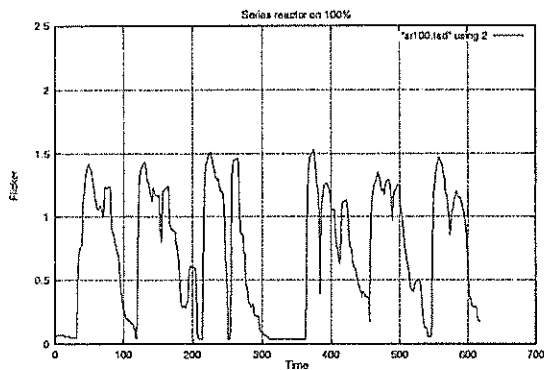
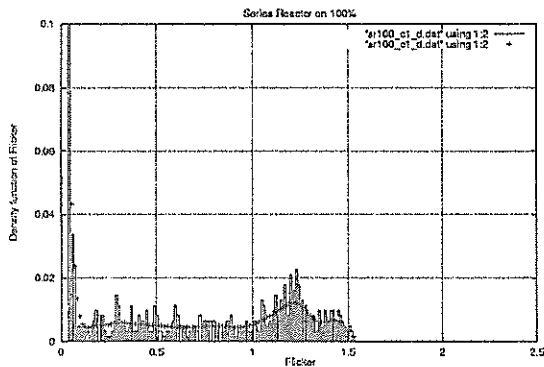


Fig. 10. Smoothed density function comparison of charging systems

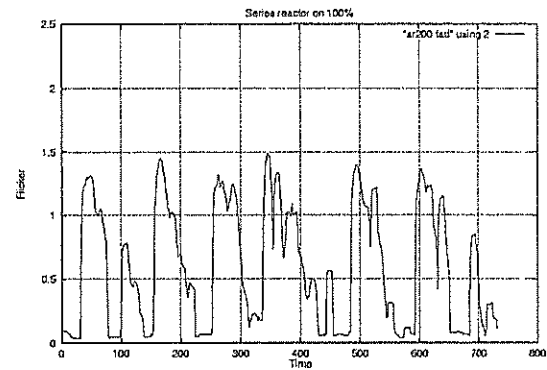


(a) Flicker vs. Time plot

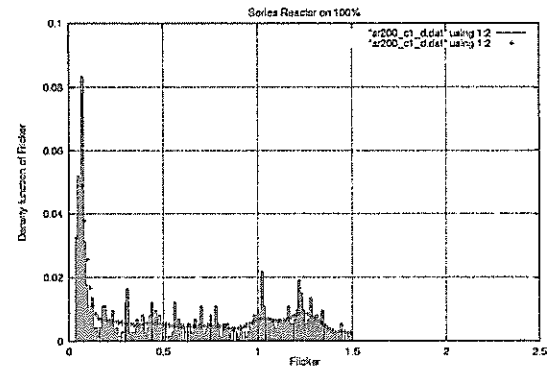


(b) Flicker density function

Fig. 5. Series Reactor (SR) 100 %



(a) Flicker vs. Time plot



(b) Flicker density function

Fig. 6. Series Reactor (SR) 100 %

basic procedure followed were to determine the density function of the flicker and to apply a smoothing function on the density function. Figure 8(a) shows the flicker vs. time data where charging took place with the orange peel system. Figure 8(b) shows the applied density function and smoothed density function where charging took place with the orange peel system. Figure 9(a) shows the flicker vs. time data where charging took place with the clamp shell system. Figure 9(b) shows the applied density function and smoothed density function where charging took place with the clamp shell system. Figure 10 shows a comparison of the orange peel and clamp shell charging systems. The density function of the flicker values for the orange peel charging system is more densely populated around higher flicker values than the clamp shell charging system.

VII. CONCLUSION

As expected the change in the Series Reactor caused a change in flicker emission. The reduction in impedance causes a reduction in current stability and network fault level S_{ccn} . From the decrease in short circuit MVA of the network seen from the

PCC the increase in flicker emission can be expected.

The findings of the charging system are not yet conclusive. The comparison of the charging systems must have the same basket order. A comparison of the charging systems where the orange peel and clamp shell systems are both used as the first basket in the melt has to be done.

By using the probability density function, it is possible to see the characteristics of a arc furnace. The difference between system parameter correlated to flicker can clearly be shown using the probability density function.

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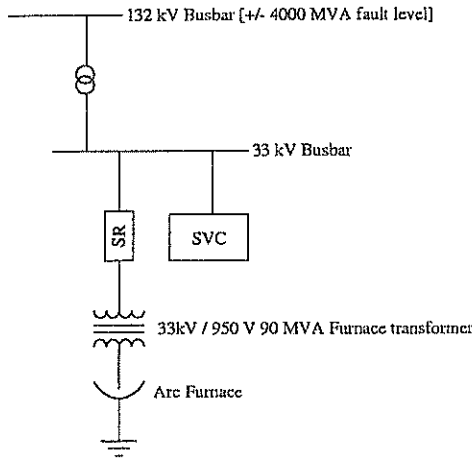


Fig. 3. One line diagram

with some statistical analysis. Where the statistical methods considered can be:

- The probability density function
- Maximum, 99%, 95% and average flicker values
- The power density spectrum

For any results to be acceptable a sufficient number of data points are needed. The probability density function gives a good indication of the characteristic of the system that is measured and is used as a base for comparison where system parameter changes are to be compared.

The probability density function for discrete random variables is given by[5]:

$$f_x(x) = \sum_{i=1}^N P(x_i) \delta(x - x_i) \quad (4)$$

where $P(x_i)$ is the probability associated with the event x_i and $\delta(x - x_i)$ is the impulse function at x_i .

Properties of the density function:

$$0 \leq f_x(x) \text{ all } x \quad (5)$$

$$\int_{-\infty}^{\infty} f_x(x) dx = 1 \quad (6)$$

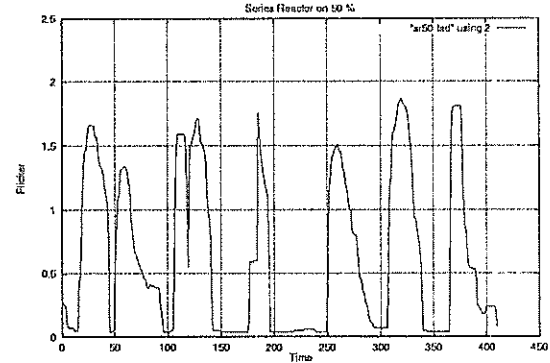
$$F_X(x) = \int_{-\infty}^x f_x(\zeta) d\zeta \quad (7)$$

$$P\{x_1 < X \leq x_2\} = \int_{x_1}^{x_2} f_x(x) dx \quad (8)$$

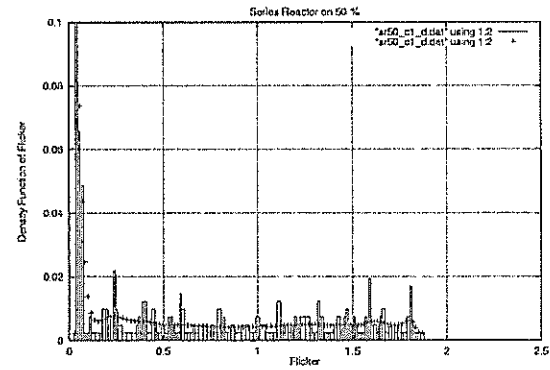
VI. RESULTS

Results showing the influence of the series reactor on flicker are shown. Six melt cycles with the SR on 50 % are compared with the same amount of melt cycles with SR on 100 %. For statistical purposes, another six melt cycles with SR on 100 % are used. This gives a total of three graphs used in the comparison. The basic procedure followed were to determine the density function of the flicker and to apply a smoothing function on the density function. Figure 4(a) shows the flicker vs. time data

for SR on 50%. Figure 4(b) shows the applied density function and smoothed density function with for SR 50%. Figure 5(a) and 6(a) shows flicker vs. time data for SR on 100%. Figure 5(b) and 6(b) shows the applied density functions and smoothed density functions with for SR 100%. The smoothed functions are then compared in figure 7. When the SR is on 50 %, higher flicker values are experienced in comparison to the SR on 100 %. The effect of the series reactor is to limit the flicker values below a certain point, in this case it is 1.5.



(a) Flicker vs. Time plot



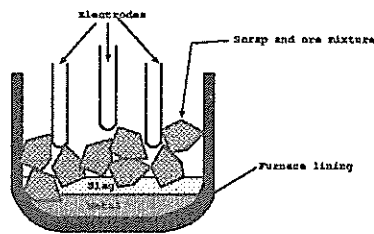
(b) Flicker density function

Fig. 4. Series Reactor (SR) 50 %

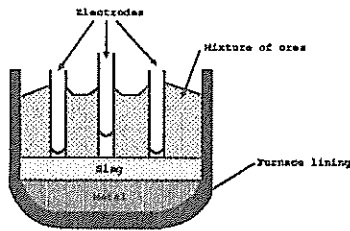
Results showing the influence of the charging system on flicker are shown. Twelve melt cycles where charging took place with the orange peel system are compared with the same amount of melt cycles where charging took place with the clamp shell system. What is important to note is that usually two baskets are used in each melt cycle. By considering only the part of the melt cycle relevant to the specific charging system a comparison is made. In this case the orange peel system is always the first basket and the clamp shell system is always the second basket. Further investigation where the basket orders are similar will be investigated. The

TABLE II
ARC FURNACE TECHNOLOGY

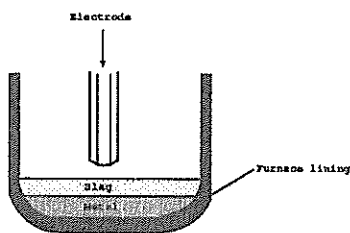
Section	Technology
Non electrical power input	Oxy-fuel burners, Oxygen lancing, CO post combustion
Electrical	Arc control, arc stability, flicker, power factor, electrode consumption
Compensation equipment	SVC, series reactor, series condenser
Other	Foamy slag practice, EAF bottom stirring, modern furnace hydraulics, scrap preheating



(a) AC open arc furnace



(b) AC submerged arc furnace



(c) DC furnace

Fig. 1. The 3 basic furnace types

IV. MEASUREMENTS

Measurements were done on a 90MVA open arc furnace with SVC and SR present. See Figure 3 for a one line diagram. The parameters considered in our field measurements to correlate with flicker are:

- Series Reactor setting:
Using the PCC at reference point, changing the value of the Series Reactor will influence the line impedance. By decreasing the SR, a increase in flicker emission is expected. This can be seen when looking at equation 1. A decrease in S_{ccn} will cause a increase in P_{st}
- Charging system:
For a scrap furnace the charging system can influence the distribution of scrap in the furnace (see figure 2 for a description of the charging systems). With a clamp shell system the scrap distribution is not even, the scrap falls out of the basket in 2 stages. A initial stage where the scrap gradually starts falling, and a final stage where the rest of the remaining scrap

falls into the furnace. With an orange peel system the distribution of scrap in the furnace is more even. The scrap falls out of the basket in one stage, everything at once. A reduction of flicker is expected for the orange peel system because the scrap has a more even distribution in the furnace, this will improve arc stability.

- Density of the scrap in the basket
The density and scrap distribution will determine the melt down time and arc stability.
- SVC (activated/deactivated)
The influence of the SVC can be determined from equation 2. Once the influence of the other system parameters has been determined, correlation between system parameters and flicker can be determined.

In most practical furnace systems only a few parameters can be measured. Small changes to furnace operation parameters are limited because of production considerations.

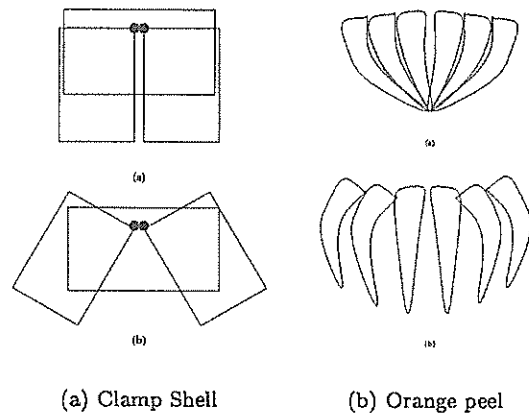


Fig. 2. Orange peel and Clamp shell charging systems

V. INVESTIGATION

The influence of parameter variations on flicker can be investigated in many ways. Usually a visualization of the flicker vs. time data together

An investigation into the influence of arc furnace system parameters on flicker emission

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Abstract- Using statistical analysis the flicker of an arc furnace is investigated to determine the influence of Series Reactance and two different charging systems. Several measurements on a 90 MVA open arc furnace are discussed

Keywords - Flicker, arc furnace, arc furnace technology

I. INTRODUCTION

The increase of industrial loads and the increase in power demand from more low voltage customers makes power quality a modern day problem [1]. Flicker is caused by voltage oscillations in the 1 to 35 Hz range and is usually experienced in lighting equipment. The industrial loads causing flicker, range from arc furnaces to steel and lumber mills. The focus of this study is on arc furnaces and the influence of system parameters on flicker [2].

II. FLICKER PREDICTION TECHNIQUES

In South Africa, the UIE/IEC flicker meters and flicker prediction techniques are used.

A flicker emission coefficient P_{st} is determined. This gives an indication of the level of flicker generated by the arc furnace. It was recognised that infrequent high values of P_{st} will not really be detrimental to voltage quality and that the value $P_{st99\%}$, i.e. the value having a probability of 1% of being exceeded on a significant measurement period is a better criterion than the maximum P_{stM} [3] [4].

The recommended prediction formula for arc furnace flicker at the point of common coupling (PCC) is (with the symbols given in table I)

$$P_{st99\%} = K_{st} \frac{S_{ccf}}{S_{ccn}} \quad (1)$$

If a SVC is present in the system, the formula becomes

$$P_{st99\%} = K_{st} \frac{S_{ccf}}{S_{ccn}} \frac{1}{R_{comp}} C_{HV/LV} \quad (2)$$

with

$$R_{SVC} = 1 + 0.75 \frac{S_{SVC}}{S_f} \quad (3)$$

for the usual range of $\frac{S_{SVC}}{S_f} < 2$.

The flicker reduction effect of a series reactor (SR) may be predicted from the expected decrease in S_{ccf} (for example 20 %) [3]

If a synchronous condenser (SC) is used, its flicker reduction effect may be predicted from the expected increase in S_{ccn} at its point of connection.[3]

The use of DC furnaces are increasing rapidly. Their flicker emission compared to an AC furnace is $\frac{1}{2} AC_{furnace(without SR)}$ and $\frac{3}{4} AC_{furnace(with SR)}$ [3].

TABLE I

DESCRIPTION OF FLICKER PREDICTION PARAMETERS

Parameter	Description
$P_{st99\%}$	Predicted flicker emission
S_{ccf}	Short circuit level of the arc furnace at the PCC
S_{ccn}	Fault level of the network at the PCC
K_{st}	Characteristic emission coefficient for P_{st} , ranging from 48 to 85 with a mean value of 60
R_{comp}	Flicker reduction factor
$C_{HV/LV}$	Attenuation coefficient
SVC	Static Var Compensator
SR	Series Reactor
SC	Series Condenser
S_{SVC}	MVA rating of SVC
S_f	MVA rating of furnace

III. ARC FURNACE TECHNOLOGY

Both AC and DC technology are used for arc furnaces in industrial applications. With basically three types of furnaces (See figure 1) AC open arc, AC submerged arc and DC furnaces.

Table II gives a furnace technology overview. These parameters play an important role in the choice of furnace. The choice of compensation equipment will in turn be determined by flicker, power factor and harmonics.

$$\frac{\arg(V(1)) - \arg(I(1))}{2 \cdot \pi} \cdot 360^\circ = 12.253^\circ$$

To estimate the possible error that could occur if the fundamental frequency should drift by 4%, the following calculation was done.

$$\left(\frac{\arg(V(1.02)) - \arg(I(1.02))}{2 \cdot \pi} \right) - \left(\frac{\arg(V(0.98)) - \arg(I(0.93))}{2 \cdot \pi} \right) \cdot 360^\circ = 5.0807^\circ$$

This value of 5.0807° is less than the previously calculated value of 6.3° .

4 Conclusion

The use of a DFT to calculate the phase angle of non-sinusoidal waveform is a viable alternative. It also has the added advantage of being easily implementable in a low cost microcontroller because of the fact that it requires a small number of calculations. The calculations can also be done as the samples arrive, minimizing idle time and using the microcontroller more effectively. The accuracy is acceptable when the leakage effect is minimized, and harmonics are not dominant.

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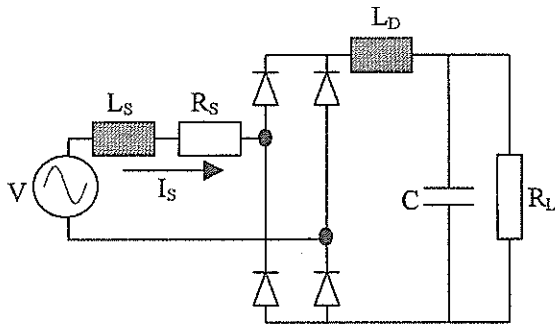


Figure 10 : 1 Phase Diode Bridge Rectifier
The inductance L_s was chosen big to give a lagging power factor. Figure 11 shows the source current I_s and the source voltage V .

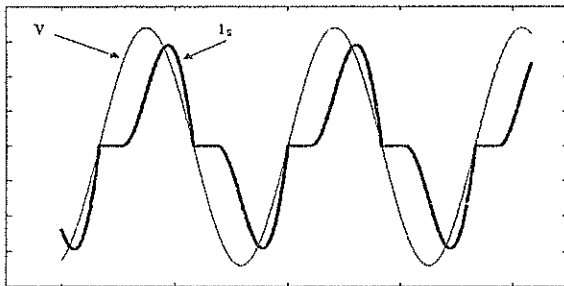


Figure 11: Voltage and current as seen from the source.

Figure 12 shows a single period of the current and voltage. It is clear that the current contains harmonic components and is phase shifted due to the lagging power factor of the load.

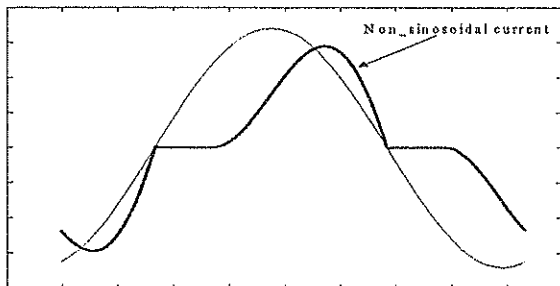


Figure 12 : A single cycle of voltage and current captured in the measuring window

By taking samples of the waveforms and performing the DFT on it, it will be possible to determine the phase shift between the fundamental components of the load voltage and currents.

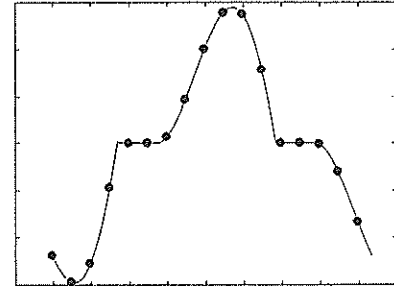


Figure 13: Sampling points for current I_s (17 points)

Figure 13 shows the sampling points for the current I_s , $i(n)$. Another advantage of the DFT is that it is not necessary to choose number of sampling points equal to the power of 2 as with the FFT.

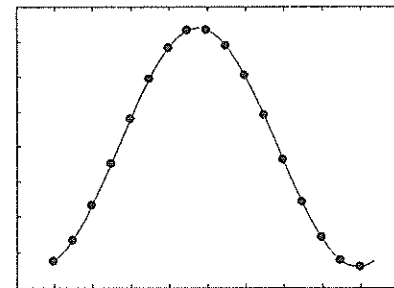


Figure 14: Sampling points for the source voltage V (17 points)

Figure 14 shows the 17 voltage sampling points ($V(n)$). The voltage V is a pure sinusoid, because the data points was generated in a Pspice simulation program. The arbitrary phase of the voltage in Figure 14 is deliberate to show that this technique will work, no matter where the sampling starts. The DFT calculations

$$V(1) = \sum_{n=0}^{17-1} V(n) \cdot e^{-j \frac{2\pi n}{17}}$$

$$I(1) = \sum_{n=0}^{17-1} I(n) \cdot e^{-j \frac{2\pi n}{17}}$$

Using Matlab the follow Phase angle was calculated in degrees

sampling. More specifically the magnitude of $X(k)$ changes by a factor of $\|\Phi(\delta)\|$ and a value of $\arg(\Phi(\delta))$ is added to the phase angle of $X(k)$. Thus we only have to look at the function $\Phi(\delta)$ to determine the changes in $X(k)$

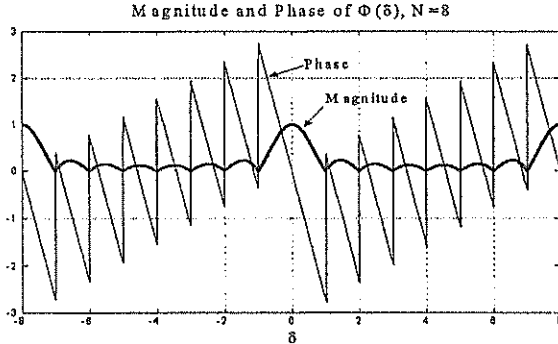


Figure 9 : Reduced interpolation function $\Phi(\delta)$

Figure 9 shows the form of the so called interpolation function, for $N=8$. The magnitude of this function bears a lot of resemblance to the continuous Fourier transform. This is because it interpolates the DFT values to the CFT values, and gives the CFT its characteristic shape.

As an example of calculating the error for frequency drift look at the following chosen scenario. Lets assume the ground frequency of 50Hz can vary from 49Hz to 51Hz. Percentage wise it corresponds to a 2% variation either way (i.e. total variation of 4%). If we were to perform the DFT to extract the phase information of this fundamental, the measured phase would vary

$$\arg(\Phi(0.02)) - \arg(\Phi(-0.02)) = 0.11 \text{ rad} = 6.3^\circ$$

This error is independent of the amplitude of the measured waveform, since $\Phi(\delta)$ is independent of the amplitude.

The question that remains is if this answer is still valid or good enough in the presence of harmonics. To simplify the problem, we assume that only the 3rd, 5th and 7th harmonics are present in the signal.

The following simplification can be made for $0 < \delta < 0.5$,

$$\left| \frac{\sin(\pi \cdot (p - k + \delta))}{N \cdot \sin\left(\frac{\pi \cdot (p - k + \delta)}{N}\right)} \right| \leq \left| \frac{\delta}{(k - p + \delta)} \right| \cdot 1.1 \quad \dots (19)$$

In Equation (15) consist out of two parts. The first exponential part specifies a unity vector in a certain direction. The second part, which can be replaced by equation (98), specifies the magnitude. Equation (14) can be rewritten as follows to include the effect of the chosen harmonics

$$\begin{aligned} X(1+\delta) &\approx X(1) \cdot \Phi(\delta) \\ &+ X(3) \cdot \exp\left[-j \cdot \frac{\pi \cdot (N-1)}{N} \cdot (\delta-2)\right] \cdot \left|\frac{\delta}{(\delta-2)}\right| \cdot 1.1 \\ &+ X(5) \cdot \exp\left[-j \cdot \frac{\pi \cdot (N-1)}{N} \cdot (\delta-4)\right] \cdot \left|\frac{\delta}{(\delta-4)}\right| \cdot 1.1 \\ &+ X(7) \cdot \exp\left[-j \cdot \frac{\pi \cdot (N-1)}{N} \cdot (\delta-6)\right] \cdot \left|\frac{\delta}{(\delta-6)}\right| \cdot 1.1 \\ &\dots (20) \end{aligned}$$

Substituting typical values for the harmonics

$$\begin{aligned} X(1+\delta) &\leq X(1) \cdot \Phi(\delta) \\ &+ (0.75 + 0.36 + 0.1) \cdot |X(1)| \cdot \left|\frac{\delta}{(\delta-2)}\right| \cdot 1.1 \angle \gamma \\ &\dots (21) \end{aligned}$$

Where γ is chosen to be the angle that would give the biggest error.

$$\begin{aligned} X(1+\delta) &\approx X(1) \cdot \Phi(\delta) + 1.33 \cdot |X(1)| \cdot \left|\frac{\delta}{(\delta-2)}\right| \angle \gamma \\ &\dots (22) \end{aligned}$$

for our previous example we again choose $\delta=0.02$

$$\begin{aligned} X(1+\delta) &\approx X(1) \cdot \Phi(\delta) + (-0.013) \cdot |X(1)| \angle \gamma \\ &\approx X(1) \cdot \Phi(\delta) \\ &\dots (23) \end{aligned}$$

The Harmonics have less than 1% influence and are therefore not significant. It must be noted though that the harmonic current magnitudes become very big, they will in fact cause a greater error.

Example

The following example considers a nonlinear load consisting of a single-phase diode bridge rectifier shown in Figure 10.

$$X(1) = \sum_{n=0}^{N-1} x(n) \cdot e^{-j \frac{2\pi \cdot n}{N}} \quad \dots (10)$$

This transform is performed on the sampled current and voltage waveforms. The relative phase angle is then easily calculated by subtracting the phase angles which is obtained from the respective transforms.

Frequency leakage in more depth

For a signal to be periodic it must be composed of the sum of pure sinusoidal signals that all have frequencies that are integer multiples of some base frequency [3],[4]. If there are to be no frequency leakage effects for a 20ms base window, the following conditions must be adhered to

- i) Lowest frequency in signal must be 50Hz
- ii) All other frequencies must be integer multiples of 50Hz

Since the current can be assumed to be periodic in the slow changing applications for which a SVC is typically used, minimal frequency leakage should occur. More specifically, if the two thyristors that control the current through one reactor, are fired symmetrically, then only uneven (integer) harmonics will be generated by the resulting current waveform. [5, 6]

If the sampling window duration should deviate from 20ms or the fundamental frequency should not be exactly 50Hz, the frequency leakage effect could become visible again, and influence the accuracy of the measured and transformed values. Since we are only interested in one specific point, we only consider the effect it will have on the accuracy of our transformed value $X(1)$. What does in fact happen when the window length deviates is that more or less than an integer cycle is sampled. For example, if ideally one cycle is to be sampled, and the sampling window duration increases, then 1.1 cycles will be sampled. This will cause a shift of the main lobe to the position of 1.1 and the DFT at $X(1)$ will no longer reflect an exact representation, i.e. there will be an error in the phase and amplitude obtained. To determine the magnitude of the error consider an equivalent problem. In this problem it is assumed that exactly an integer amount of cycles are sampled in the sampling window. Then determining the values for $X(k+\delta)$, where k is an integer and δ is a fraction, it will be equivalent to shifting the main lobe left or right by an amount δ , and then determining $X(k)$. This will give the same result as increasing the sampling window δ percent. Setting out to find an expression for $X(k+\delta)$, we notice the following. The original samples, $x(n)$ can be

determined from $X(k)$ by the inverse Fourier transform and $X(e^{j\omega})$ from $x(n)$ by the Fourier transform. Therefore a way of determining $X(e^{j\omega})$ from $X(k)$ must exist. It can be shown that[4]

$$X(e^{j\omega}) = \sum_{p=0}^{N-1} X(p) \cdot \Phi\left(\omega - \frac{2\pi}{N} \cdot p\right) \quad \dots (11)$$

where

$$\Phi(\omega) = \exp\left(-j \frac{(N-1)}{2} \cdot \omega\right) \cdot \frac{\sin\left(\frac{\omega \cdot N}{2}\right)}{N \cdot \sin\left(\frac{\omega}{2}\right)} \quad \dots (12)$$

The equation (11) is called the DFT interpolation formula, and the $\Phi(\omega)$ is the interpolation function. It should be emphasized that $x(n)$ must be of finite duration for this result to hold. If we substitute into (11)

$$\omega = \frac{2\pi}{N} (k + \delta) \quad \dots (13)$$

the equation becomes

$$X(k + \delta) = \sum_{p=0}^{N-1} X(p) \cdot \Phi\left(\frac{2\pi}{N} (k + \delta - p)\right) \quad \dots (14)$$

and $\Phi\left(\frac{2\pi}{N} (k + \delta - p)\right)$ can be expressed as

$$\Phi(k - p + \delta) = \exp\left(-j \cdot \frac{(N-1) \cdot \pi}{N} (k - p + \delta)\right) \cdot \frac{\sin(\pi \cdot (k - p + \delta))}{N \cdot \sin\left(\frac{\pi}{N} (k - p + \delta)\right)} \quad \dots (15)$$

And (14) becomes

$$X(k + \delta) = \sum_{p=0}^{N-1} X(p) \cdot \Phi(p - k + \delta) \quad \dots (16)$$

With $\delta=0$ we imply all frequencies present in the signal are integer multiples of the fundamental frequency. Assume that only a single frequency is present in the sampled window, namely the frequency at $X(k)$. All the other points in the DFT are then zero, except to $X(k)$. Thus the equation further reduces to

$$X(k + \delta) = X(k) \cdot \Phi(\delta) \quad \dots (17)$$

and because $X(k)$ and $\Phi(\delta)$ are complex values we can write it in term of phasor notation.

$$X(k + \delta) = \|X(k)\| \cdot \|\Phi(\delta)\| \angle [\arg(X(k)) + \arg(\Phi(\delta))] \quad \dots (18)$$

This expression shows how $X(k)$ changes if the sampling window size changes with a factor δ . For example $\delta=0$ signifies the zero relative change of the sampling window size and $\delta=0.5$ signifies a 50% change in

in the N sampled value, so-called frequency leakage takes place, and $X(k)$ no longer have the same meaning. The meaning of the frequency leakage effect and how it effects the accuracy of the transformation is discussed later.

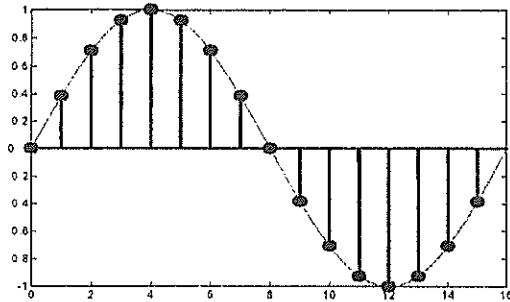


Figure 5: 1 period in base window (16 samples)

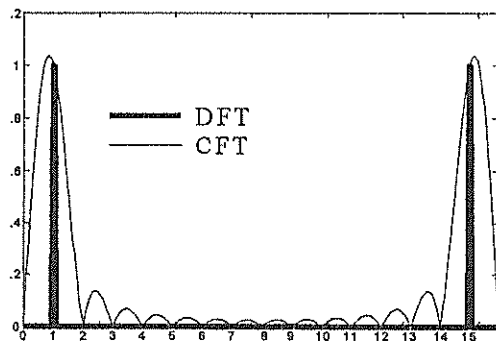


Figure 6: DFT & CFT of 1 period sampled

Figure 6 shows the CFT and DFT of a sampled sinus-waveform of which exactly one cycle fits into the base window. The main lobe of the FT is in the region of $k=1$, i.e. where $\omega = \frac{2 \cdot \pi \cdot 1}{N}$. The DFT shows the points where the CFT is sampled, and in this case gives a clear representation of the frequency content of the signal.

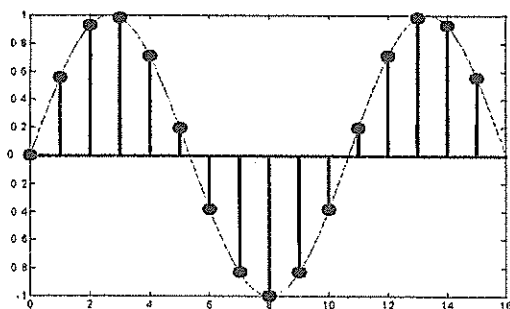


Figure 7: 1.5 periods in base window (16 samples)

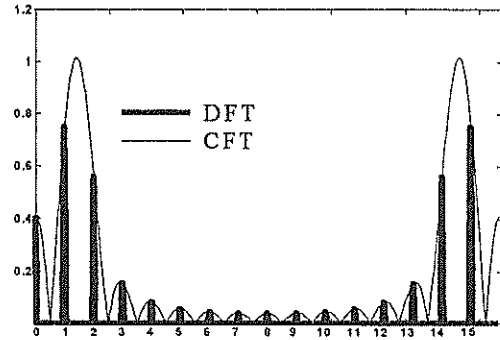


Figure 8: DFT & CFT of 1.5 periods sampled

In Figure 7 the signal completes 1.5 cycles in the allowed sampling points. Now the main lobe of the CFT in Figure 8 is centered around $k=1.5$ to reflect the 1.5 cycles, but because the DFT samples the CFT at fixed integer intervals, the resulting DFT shows a lot more harmonics present in the system. This illustrates the frequency leakage effect. Ways to minimize this effect is to take more sampling points or the use of windowing functions. Windowing functions multiply each sampled point by a certain weighting factor. Keep in mind that windowing functions could give misleading results, if the frequencies present in the system are not far enough apart. This is because a windowing function changes the width and amplitudes of the lobes in the CFT. [3]

Fundamental frequency

Since the SVC can only compensate for the fundamental component of the current, we need only do a transform to determine one point, namely for the 50Hz component. Normally a fast Fourier transform would be the first choice, but since we need only calculate one point, a normal discrete Fourier transform is in fact the better option. The DFT does fewer calculations compared to a FFT if the calculation of only one point is desired. Another advantage of the DFT is that the calculations can be done as the A/D samples arrive, as opposed to the FFT routine that only performs the transform after all the samples are taken.

If we choose the sampling (base) window duration as 20ms, $X(1)$ will carry the information of the (50Hz) fundamental component. Next we choose a sampling frequency of 1.6 kHz with the assumption that 800Hz will be the highest frequency present in the measured signal (an anti-aliasing filter should be used in practice). Though this will give a resolution of 32 points, we are only interested in $X(1)$. For $X(1)$ the full DFT calculation reduces to N complex multiplication and additions.

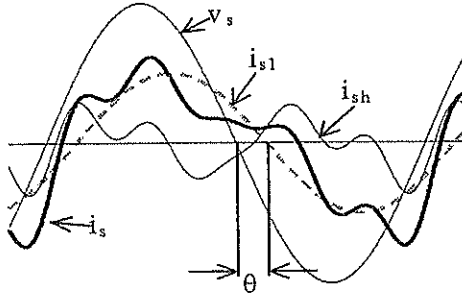


Figure 4: Line current distortion

If we assume a pure sinusoidal supply voltage v_s , then i_s may be expressed as the sum of its Fourier (harmonic) components (dc component assumed to be zero)

$$i_s(t) = i_{s1}(t) + \sum_{h \neq 1} i_{sh}(t) \quad \dots (2)$$

Where i_{s1} is the fundamental (line frequency f_1) component, i_{sh} the component at the harmonic frequency $f_h = hf_1$ and θ the phase angle between the assumed sinusoid voltage v_s and the current I_s . The average power is given by

$$P = V_s \cdot I_s \cdot \cos\theta \quad \dots (3)$$

Where I_s and V_s is the RMS value of i_s and v_s respectively. The current components at harmonic frequencies do not contribute to the average real power drawn from the sinusoidal voltage source v_s . The apparent power is:

$$S = V_s \cdot I_s \quad \dots (4)$$

The power factor then per definition is

$$PF = \frac{P}{S} = \frac{V_s \cdot I_{s1} \cdot \cos\theta}{V_s \cdot I_s} = \frac{I_{s1}}{I_s} \cdot \cos\theta \quad \dots (5)$$

With I_{s1} the RMS value is i_{s1} . The displacement power factor is defined as :

$$DPF = \cos\theta \quad \dots (6)$$

2 Determining the DPF

In order to control a SVC it is useful to determine the displacement power factor of the load (as seen from the source). This serves as an input value for the actual control scheme. Many methods exist for controlling a SVC. Among these are

- Using a low pass filter to allow only the fundamental frequency to pass through and then measure the time difference between the zero crossings of the voltage and current waveforms. This gives a good indication of the phase angle between the voltage and current waveform. A drawback is the phase shift caused by the filters (time delay).

- Calculation of the effective susceptance of the load by sampling the phase currents. This is done by sampling the phase currents and then comparing it to a stored cosine waveform, by phase shifting the samples until a best fit is found. (correlation). [2]
- Optimal load compensation by calculating a triggering angle which minimizes the RMS current value of the total current (load + compensator). It has the drawback of being mathematically intensive. [6]

2.1 Fourier transform methods

The Fourier transform enables us to determine the amplitude and phase properties of the frequency components in a measured signal. With the advent of DSP's, computers and relatively accurate A/D equipment, it has become increasingly easier to perform the discrete Fourier transformation (DFT) on a set of sampled data points. This set of data points is representative of a sampled analog signal or simulated values. If the Fourier transform is performed on these measured sampled points, the phase and amplitude of the signal can be read from the transformed data, which can later be used in the control scheme of the SVC. The DFT makes it possible to estimate the fundamental component of the non-sinusoidal waveform. This can then be used to determine the DPF. To fully understand all the effects of the Fourier transform it is necessary look at how it works.

The sampled values of a continuous time signal $x(t)$ is represented by $x(n)$, where $x(n) = x(t)|_{t=nT}$ and T is the sampling period. The continuous Fourier transform (CFT) of $x(n)$ which is zero outside $n=0, 1, 2, \dots, N-1$, is given by

$$X(e^{j\omega}) = \sum_{n=-\infty}^{\infty} x(n) \cdot e^{-j\omega n} = \sum_{n=0}^{N-1} x(n) \cdot e^{-j\omega n} \quad \dots (7)$$

It is worth noting that $X(e^{j\omega})$ is a continuous function of ω , and periodic for a period of $\omega=2\pi$. The discrete Fourier transform of N sampled points $x(n)$, is given by

$$X(k) = \sum_{n=0}^{N-1} x(n) \cdot e^{-j\frac{2\pi \cdot k \cdot n}{N}} \quad \dots (8)$$

From this equation it can be seen that the DFT is a sampled version of the full CFT,

$$X(k) = X(e^{j\omega}) \Big|_{\omega = \frac{2\pi \cdot k}{N}} \quad \dots (9)$$

The value $X(k)$ ideally represents the phase and angle of the sinusoid that has completed exactly k integer cycles in the N sampled values. In cases where a sinusoid is sampled that does not have integer multiples of its cycles

Assessing the viability of using a Discrete Fourier Transform to determine the displacement power factor.

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Abstract: In the control of a static var compensator, it is necessary to know the displacement power factor in order to compensate for the reactive part of the load. If the current waveform is distorted this becomes difficult and filtering out techniques is required. In this paper the displacement power factor is found by means of a discrete Fourier transform on a cycle by cycle basis. The transform is optimized for this application

Keywords : Static VAR compensator, discrete Fourier transform, displacement power factor

1 INTRODUCTION

The purpose of a Static VAR compensator (SVC) is to minimize the amount of reactive current supplied to the load by the source. That current, which is necessary for the proper operation of the load, is then supplied by the SVC by effectively placing the appropriate inductive or capacitive component in parallel with the load. The SVC is then said to supply the reactive power needed by the load.

The SVC to which will be referred to in this paper employs a thyristor controlled reactor in parallel with a fixed capacitor bank (TCRFC). The TCRFC works on the principle of placing a variable reactor in parallel with a capacitor. This variable reactor is realized by using thyristors to control the current flow through a fixed reactor.

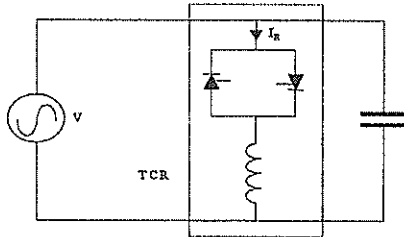


Figure 19: TCR in parallel with the capacitor bank

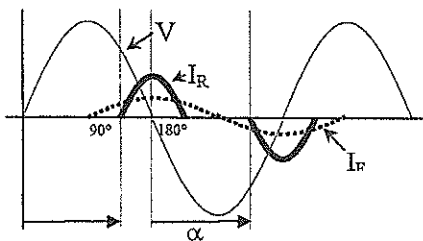


Figure 2: Current flow in reactor for a firing angle(α)

Figure 2 shows the current flowing in the reactor(I_R) if the thyristors are fired at an angle α . I_R consists out of a 50Hz fundamental component I_F as well as other harmonic components. By varying the firing angle the amount of current that flows through the reactor is controlled, i.e. the amplitude of the fundamental current I_F . The apparent susceptance of the reactor as function of firing angle α is given by [1]

$$B(\sigma) = \frac{\sigma - \sin \sigma}{\pi \cdot X_L} \quad \dots (1)$$

$$\text{Where } \sigma = 2 \cdot (\pi - \alpha)$$

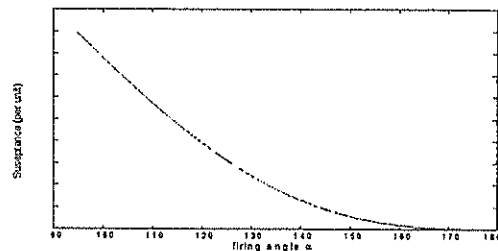


Figure 3: Apparent susceptance of TCR as function of firing angle α

1.1 Displacement power factor

Figure 4 shows a line current i_s drawn from the utility by the load, that deviates significantly from a sinusoidal voltage waveform.

4. Conclusion

Based on the results obtained, one can conclude that AVRs are important components to include in systems especially where the loads are modeled as constant power. The system without the AVR installed resulted with system voltages after the 50 msec disturbance lower than when with the AVR.

Because the constant power load model gives rise to a worst case scenario, one has to have in their model the controllers that would bring back the system voltage to more realistic values.

In both systems, as suggested before, eigenvalue analysis can be performed to determine the stability of the system. Furthermore, in cases of much bigger systems, remedial actions might need to be taken. With the use of participation factors, they can determine where the weak busbars are located and apply the appropriate remedial action. Unfortunately, one can not choose appropriate remedial action based on values obtained from the participation factors.

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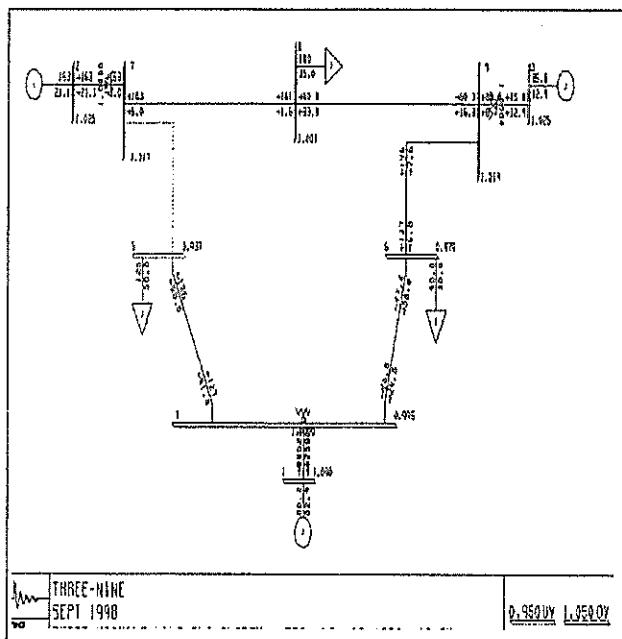


Fig. 3 Line 5-7 out for 50 msecs

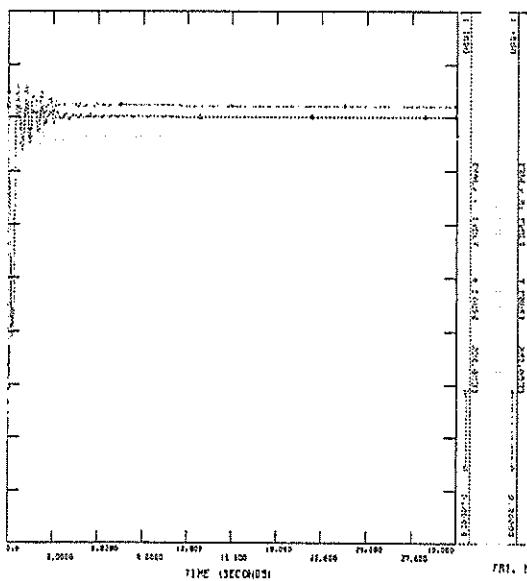


Fig. 4 System without AVR

The same disturbance was applied in the dynamic simulations on the same system with and without Automatic Voltage Regulators. Fig. 4 and Fig. 5 show the outputs respectively.

Comparing both Fig. 4 and Fig. 5, one will notice that the voltage level at which the bus 5 settles is low. While with an AVR in the system, the voltages are raised to acceptable values.

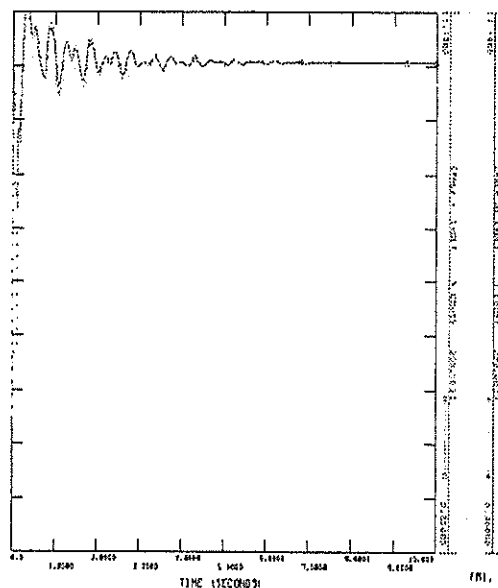


Fig. 5 System with AVR

Extension

An extension to what has already been done is the possible eigenvalue analysis. This performed on both systems with and without an AVR will result in slightly differences in values. This would be due to the change in the system characteristics (the addition of AVR).

2. Basic Mathematics

The free response of the system can be determined from the eigenvalues of the A matrix. [4,5,6]

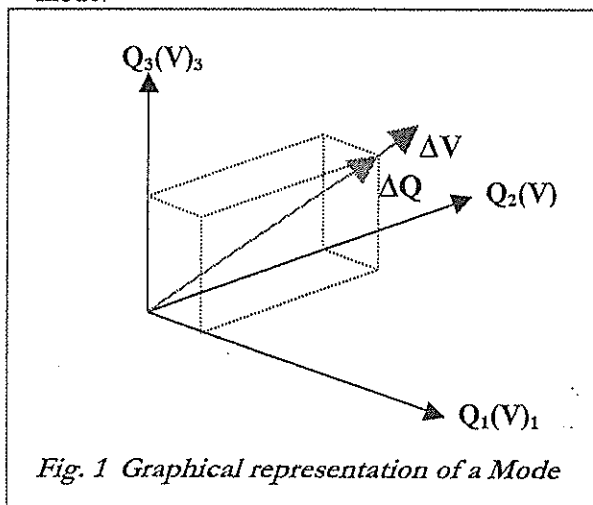
Eigenvalues of the matrix A are the roots of the characteristic equations [7]

$$\dot{x} = Ax + Bu$$

The number of eigenvalues is equal to the number of states in the system. For a stable system, all the eigenvalues have negative real parts while the magnitude of the imaginary part will tell how oscillatory the mode is.

A mode is described as a way of disturbing the system by imposing reactive power injections (in a give direction) which results in proportional bus voltage change (in the same direction). For graphical representation of a mode is in Fig. 1.

Using the eigenvalues, the right and left eigenvector is obtained for each eigenvalue. Mathematically manipulated, one obtains what is known as the participation factor. Participation factors is defined as a measure of the relative participation of a state variable in a certain mode with the use of both the right and the left eigenvectors of a mode.



3. Example and Results

The three machine nine bus is represented graphically using PSS/E in Fig.2. The load flow was simulated to obtain load equilibria.

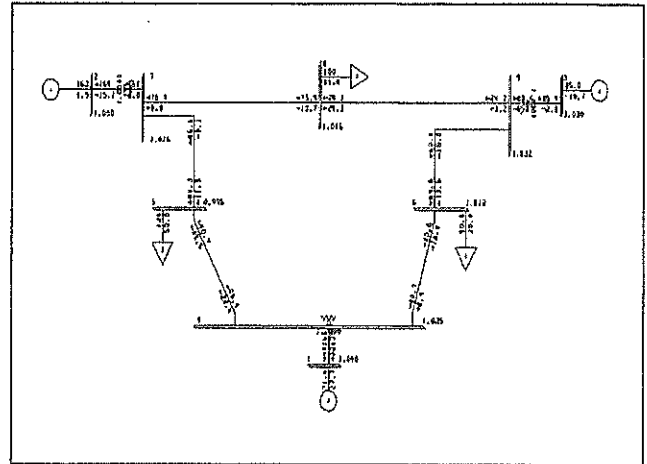


Fig. 2 Load Flow of Three Machine Nine Bus System

Closely examining the systems load flow, the line between bus 5 and bus 7 (line 5-7) is the heaviest loaded line in terms of real power. This line has been chosen for the 50 msec tripping. This will lead to busbars hitting the lower voltage limit compare to the trip of the line between bus 6 and bus 9 (line 6-9) which is the biggest producer of reactive power.

Governors where not included. An AVR (Automatic Voltage Regulator) controls the output of the synchronous generator by controlling the exciter. [4]

The are many type of AVR, some electromechanical, the older one and continuously acting which are faster. In this paper a continuously acting AVR's will be considered.

In the dynamic analysis of the system, Fig. 3 shows the load flow (Static simulation) results of the system with the line 5-7 out for 50 msec's. Analysing the system, one will notice that bus bar 5 is below the lower voltage limit.

Voltage Stability in a Small System

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Abstract:

This paper treats a three machine nine bus system in the analysis of voltage problems using various analysis. The action of an Automatic voltage Regulator is analysed in this small system and extending to eigenvalue analysis.

Keywords:

Voltage Stability, Automatic Voltage Regulator (AVR), Eigenvalue analysis, Participation factor.

1. Introduction and Definitions

The structure of the Electric power systems is getting more and more complex. This brings about many problem in solving and analysing the system due to its non-linearity. The one of the major analytical problems is the voltage stability issue which is occurring more often in our modern systems due to the increased in transfer between a remote generation and a load.[1]

What is voltage stability? There are many definitions to what Voltage Stability is. According to CIGRÉ:

A power system at a given operating state is *small-disturbance voltage stable* if, following any small disturbance, voltage near loads are identical or close to the pre-disturbance. (Small-disturbance voltage stability corresponds to a related linearised

dynamic model; with eigenvalues having real parts. [2])

When the stability of the system is being investigated, it is often convenient to assume that the disturbances causing the changes disappear. The motion of the system is then free. Stability is then assured if the system returns to its identical or very close to its original pre-disturbed condition. Such behaviour can be determined in a linear system by the characteristic equation of the system. If the mathematical description of the system is in state-space form, i.e. if the system is described by a set of first-order differential equations,

In the network, an unbalance between the power input to the system and the power consumed takes place, resulting in a transient. When this transient subsides and a steady-state condition is reached, the power impact is "shared" by various synchronous machines according to their steady-state characteristics.

System responds depends mainly on the modelling of the system. For example, load modelling in a system is important. In a real electric power system, it's not easy to model correctly the actual load for it is complex. Static load model will be used. Since the system frequency is maintained within very narrow limits, in this paper, loads will be voltage dependent and not frequency. Constant power load model will be used for all three loads in the system. This will correspond to the worst case response for any system.[3]

the energy loss results between Model A and Model C will be illustrated by example.

The example is based on data obtained during the light load period of Manyatseng township near Ladybrand in the Freestate.

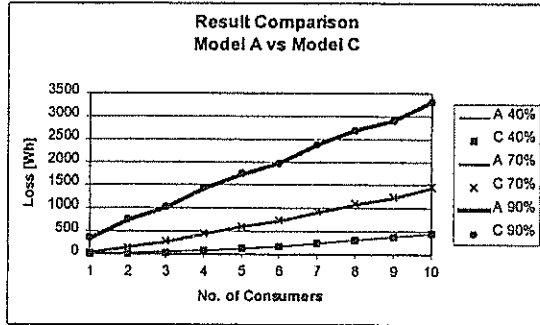


Figure 2: Comparison of energy loss results

Figure 2 shows the energy loss results for a year from 1 to 10 consumers at 40%, 70% and 90% confidence levels. The curves illustrate that, for this example, the results of the model correlate well and discrepancies cannot be observed on the graph scale.

CONCLUSION

With the models and methods developed in this papers designers of LV electrical networks have the tools to reasonably accurately assess electrical losses.

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APPENDIX – MOMENTS OF A BETA PDF

$$E(Y_i) = \frac{\alpha_i}{\alpha_i + \beta_i}$$

$$E(Y_i^2) = \frac{\alpha_i(\alpha_i + 1)}{(\alpha_i + \beta_i)(\alpha_i + \beta_i + 1)}$$

$$E(Y_i^3) = \frac{\alpha_i(\alpha_i + 1)(\alpha_i + 2)}{(\alpha_i + \beta_i)(\alpha_i + \beta_i + 1)(\alpha_i + \beta_i + 2)}$$

$$E(Y_i^4) = \frac{\alpha_i(\alpha_i + 1)(\alpha_i + 2)(\alpha_i + 3)}{(\alpha_i + \beta_i)(\alpha_i + \beta_i + 1)(\alpha_i + \beta_i + 2)(\alpha_i + \beta_i + 3)}$$

$$E(Y_i, Y_j) = \frac{\alpha_i^2}{(\alpha_i + \beta_i)^2}$$

$$E(Y_i, Y_j, Y_k) = \frac{\alpha_i^3}{(\alpha_i + \beta_i)^3}$$

$$E(Y_i, Y_j, Y_k, Y_l) = \frac{\alpha_i^4}{(\alpha_i + \beta_i)^4}$$

$$E(Y_i^2, Y_j) = \frac{\alpha_i^2(\alpha_i + 1)}{(\alpha_i + \beta_i)^2(\alpha_i + \beta_i + 1)}$$

$$E(Y_i^2, Y_j^2) = \frac{\alpha_i^2(\alpha_i + 1)^2}{(\alpha_i + \beta_i)^2(\alpha_i + \beta_i + 1)^2}$$

$$E(Y_i^3, Y_j) = \frac{\alpha_i^3(\alpha_i + 1)(\alpha_i + 2)}{(\alpha_i + \beta_i)^3(\alpha_i + \beta_i + 1)(\alpha_i + \beta_i + 2)}$$

$$E(Y_i^2, Y_j, Y_k) = \frac{\alpha_i^3(\alpha_i + 1)}{(\alpha_i + \beta_i)^3(\alpha_i + \beta_i + 1)}$$

P_{loss} in the neutral:

$$\begin{aligned} P_{loss n} &= C_b^2 R_n \left| Y_a(1+j0) + Y_b(-1/2+j\sqrt{3}/2) + Y_c(-1/2-j\sqrt{3}/2) \right|^2 \\ &= C_b^2 R_n [Y_a^2 + Y_b^2 + Y_c^2 - Y_a Y_b - Y_a Y_c - Y_b Y_c] \\ &= C_b^2 [R_p (Y_a^2 + Y_b^2 + Y_c^2) + R_n (Y_a^2 + Y_b^2 + Y_c^2 - Y_a Y_b - Y_a Y_c - Y_b Y_c)] \end{aligned} \quad \dots (10)$$

$$\begin{aligned} \text{if } R_n = k, R_p \text{ then} \\ &= C_b^2 R_p [(1+k)(Y_a^2 + Y_b^2 + Y_c^2) - k(Y_a Y_b + Y_a Y_c + Y_b Y_c)] \end{aligned} \quad \dots (11)$$

Therefore the mean or expected value for P_{loss} or $E(P_{loss})$ can be expressed as follows:

$$\mu_L = E(P_{loss}) = C_b^2 R_p [(1+k)(E(Y_a^2) + E(Y_b^2) + E(Y_c^2)) - k(E(Y_a Y_b) + E(Y_a Y_c) + E(Y_b Y_c))] \quad \dots (12)$$

The variance of P_{loss} or $VAR(P_{loss})$ can be calculated using the following expression:

$$\begin{aligned} \sigma_L^2 = VAR(P_{loss}) &= E(P_{loss}^2) - E(P_{loss})^2 \text{ and} \\ P_{loss}^2 &= C_b^4 R_p^2 [(1+k)^2 (Y_a^4 + 2Y_a^2 Y_b^2 + 2Y_a^2 Y_c^2 + Y_b^4 + 2Y_b^2 Y_c^2 + Y_c^4) + k^2 (Y_a^2 Y_b^2 + 2Y_a^2 Y_b Y_c + 2Y_a Y_b^2 Y_c + Y_a^2 Y_c^2 + 2Y_a Y_b Y_c^2 + Y_b^2 Y_c^2) - 2k(1+k)(Y_a^3 Y_b + Y_a^3 Y_c + Y_a^2 Y_b Y_c + Y_a Y_b^3 + Y_a Y_b^2 Y_c + Y_b^3 Y_c + Y_a Y_b Y_c^2 + Y_a Y_c^3 + Y_b Y_c^3)] \end{aligned} \quad \dots (13)$$

Therefore the expected value of P_{loss}^2 is:

$$\begin{aligned} E(P_{loss}^2) &= C_b^4 R_p^2 [(1+k)^2 (E(Y_a^4) + 2E(Y_a^2 Y_b^2) + 2E(Y_a^2 Y_c^2) + E(Y_b^4) + 2E(Y_b^2 Y_c^2) + E(Y_c^4)) + k^2 (E(Y_a^2 Y_b^2) + 2E(Y_a^2 Y_b Y_c) + 2E(Y_a Y_b^2 Y_c) + E(Y_a^2 Y_c^2) + 2E(Y_a Y_b Y_c^2) + E(Y_b^2 Y_c^2)) - 2k(1+k)(E(Y_a^3 Y_b) + E(Y_a^3 Y_c) + E(Y_a^2 Y_b Y_c) + E(Y_a Y_b^3) + E(Y_a Y_b^2 Y_c) + E(Y_b^3 Y_c) + E(Y_a Y_b Y_c^2) + E(Y_a Y_c^3) + E(Y_b Y_c^3))] \end{aligned} \quad \dots (14)$$

Where:

$$\begin{aligned} E(Y_a) &= m_a E(Y_i) \\ E(Y_a^2) &= m_a E(Y_i^2) + m_a(m_a-1)E(Y_j Y_k) \\ E(Y_a^3) &= m_a E(Y_i^3) + 3m_a(m_a-1)E(Y_j^2 Y_k) + m_a(m_a^2-3m_a+2)E(Y_l Y_m Y_n) \\ E(Y_a^4) &= m_a E(Y_i^4) + 4m_a(m_a-1)E(Y_j^3 Y_k) + 3m_a(m_a-1)E(Y_l^2 Y_m^2) + 6m_a(m_a^2-3m_a+2)E(Y_n^2 Y_o Y_p) + m_a(m_a^3-6m_a^2-11m_a-6)E(Y_q Y_r Y_s Y_t) \end{aligned}$$

The above expected values are only valid because the current variates are identically distributed and independent. In this case the current variates have the same Beta pdf described by parameters α_i , β_i and C_b . Expressions for the moments (expected values) of the Beta pdf are given in the appendix.

From the above equations the power loss mean (μ_L) and variance (σ_L^2) are derived. From the power loss mean and variance the loss pdf parameters can be calculated using the expressions presented in the Beta general

properties section. Due to the nature of the Beta pdf the power loss upper limit L_{max} is required to calculate Alpha (α_L) and Beta (β_L).

L_{max} is defined as the upper limit of the power loss pdf function. This occurs when all consumers at a particular node are consuming current at their upper limit, viz. C_b . This means that $Y_a = m_a$, $Y_b = m_b$ and $Y_c = m_c$ and therefore results in the following equation:

$$L_{max} = C_b^2 R_p [(1+k)(m_a^2 + m_b^2 + m_c^2) - k(m_a m_b + m_a m_c + m_b m_c)] \quad \dots (15)$$

6. MONTE CARLO SIMULATION

To ensure the analytical loss calculations are correct, the analytical results were compared with the results from a 'bootstrapped' Monte Carlo simulation (MC).

Below the cases for 3, 4 and 5 consumers on a single section network (see Figure 2) representing the three degrees of phase imbalance is presented. For this example the current pdf has parameters, $\alpha=1.93$, $\beta=17.33$ and $C_b=10$ A and the feeder resistance equals 0.1ohm with $k=0.880$ (Typical for 40m of 35mm² aerial bundled conductor.)

Table 1 contains the summary results of 30 MC simulations for 3, 4 and 5 consumers at a 90% confidence level. The similarity between the MC average and the analytical loss result confirms the validity of the method developed.

Table 1: Comparison between the MC and analytical results

No. of consumers (m)	Ploss 90% [W]		
	3	4	5
Average MC	1.16	2.08	2.66
Std Dev	0.032	0.065	0.057
%Std Dev	2.80%	3.12%	2.15%
Analytical	1.16	2.08	2.66

7. ENERGY LOSS CALCULATION

Using the loss developed calculation method, the loss Beta parametric descriptors can be calculated at each five-minute interval and evaluated at a particular confidence level. These calculated five-minute interval losses are summed over the total period for which the calculation is required. This model (Model A) is calculation intensive and requires large computer memory for long calculation sequences. It is however considered as an exhaustive benchmark for comparing other proposed models such as Model C which is much less calculation intensive. The difference in accuracy of

The load at any particular time 'slice' on a load profile has a distribution described by a set of parameters α_1 and β_1 scaling factor C_b . The load profile is therefore not a clear single line, but a range of values with central tendency and dispersion at any time.

A particular load current is dependent on the shape of the distribution function and the risk or confidence. The National Load Research raw data on which the analysis is based has a 5- minute averaging period. For LV feeders the daily load profile, using 5-minute samples, can therefore be represented by 288 (12 x 24) unique Beta pdf's. This model is however cumbersome and an alternative concise model is required.

4. BETA PDF LOAD CURRENT MODEL

A load current model must represent the load current over a period of time for energy loss calculations. The RMS value will be utilised to represent the central tendency of the load current mean over a period.

The outstanding parameter for the load profile model is its dispersion over the period. The central limit theorem states that as the population sample size increases the sample distribution approaches a Gaussian distribution regardless of the shape of the sample distribution. Generally this approximation is good if the sample size is greater than 30 [3]. Note that Gaussian distributed load currents do not imply that the loss distribution is also Gaussian as it is a function of the load current squared.

An argument therefore exists to sum or average the variances of each sample representing the load profile to obtain an indication of the dispersion. This implies a pooled estimate of the variances, which can be validated using Cochran's test [4]. This proposed current model will be referred to as Model C.

5. LOSS CALCULATION METHOD FOR BETA DISTRIBUTED LOAD CURRENTS

Figure 1 illustrates the schematic diagram on which this method is based. Let $Y_1 \dots Y_m$ be the individual, scaled stochastic load current variates associated with m consumers at a node. These variates are independent and identically distributed according to the Beta probability density function with parameters α_1 , β_1 and C_b . C_b is usually the circuit breaker size supplying the individual load, i.e. $I_i = C_b \cdot Y_i$. Then the scaled cumulative current for each phase at a particular distribution point is:

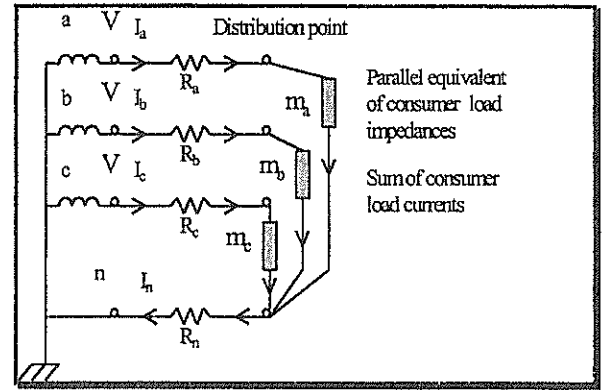


Figure 1: Schematic for single section LV feeder

$$\begin{aligned} Y_a &= (Y_1 + Y_2 + \dots + Y_{m_a}) \\ Y_b &= (Y_{m_a+1} + Y_{m_a+2} + \dots + Y_{m_a+m_b}) \\ Y_c &= (Y_{m_a+m_b+1} + Y_{m_a+m_b+2} + \dots + Y_{m_a+m_b+m_c}) \dots (5) \end{aligned}$$

m = Total no. of consumers
 m_a , m_b and m_c = no. of consumers connected to phase a, b and c respectively

The load current phasors are calculated as follows:

$$\begin{aligned} I_a &= C_b \cdot Y_a \\ I_b &= C_b \cdot Y_b \\ I_c &= C_b \cdot Y_c \dots (6) \end{aligned}$$

Only the resistive component of the line impedance generates significant losses, so for this study the line capacitive and inductive reactances are ignored. This assumption holds firm due to the following reasons:

- The load is modeled as a current load and the line capacitive and inductive reactances have no effect on its magnitude.
- No significant losses are generated in the line capacitive and inductive reactances.
- The dominant line impedance component is resistive.

The neutral current (I_n) is the sum of the load current phasors: $I_n = I_a + I_b + I_c$ where:

$$\begin{aligned} I_a &= I_a \angle 0^\circ = I_a (1 + j0) \\ I_b &= I_b \angle 120^\circ = I_b (-1/2 + j \sqrt{3}/2) \\ I_c &= I_c \angle -120^\circ = I_c (-1/2 - j \sqrt{3}/2) \dots (7) \end{aligned}$$

Total power losses (P_{loss}) for the general case will be the algebraic sum of the loss contributions of all consumers in the respective phase resistances and the vectorial sum in the neutral resistance, i.e.:

$$\begin{aligned} P_{loss} &= I_a^2 \cdot R_p + I_b^2 \cdot R_p + I_c^2 \cdot R_p + I_n^2 \cdot R_n \\ &= (I_a^2 + I_b^2 + I_c^2) \cdot R_p + |I_a + I_b + I_c|^2 \cdot R_n \dots (8) \end{aligned}$$

P_{loss} in the phases:

$$P_{loss p} = C_b^2 \cdot R_p \cdot (Y_a^2 + Y_b^2 + Y_c^2) \dots (9)$$

A PRELIMINARY INVESTIGATION INTO DOMESTIC FEEDER LOSS CALCULATION METHODS AND MODELS

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Abstract: Traditional methods of quantifying the losses in feeders use loss load factors. The method developed in this paper uses a statistical load model described by a Beta pdf. This method uses two characteristic parameters derived from load surveys. The results are compared with Monte Carlo simulations and the initial indications show that it is an acceptable approach to loss calculations in LV feeders.

Keywords: Probabilistic methods, Beta pdf, Loss calculation

1. INTRODUCTION

Developments in voltage regulation calculation require that load currents should be represented statistically. Kritzing [1] proved that the load current of a number of independent consumers at maximum demand, or any other synchronous time slice, could be represented by a **Beta probability density function**. In other words all load currents are independent and identically distributed according to the characteristic Beta probability function.

Herman [2] in his work used this result to represent the statistical voltage regulation over a feeder at load maximum demand. A feeder's maximum voltage drop can therefore be represented by a Beta probability density function (pdf) at the instant in time at which maximum demand occurs.

The aim of this paper is to propose a method to calculate energy losses in Watt-hours [Wh] on LV feeder systems due to **Beta distributed load currents**.

The first step in calculating energy losses is to calculate a power loss corresponding to a load current and therefore at a particular time 'slice'.

Once this is accomplished the energy loss can be calculated. Due to the nature of the calculation of energy losses, power loss calculations over a period of time are required for summation. The second problem therefore is to represent the losses or possibly the load current over the period for which energy losses need to be calculated. This period can typically be a day, week, month, season or year.

1. GENERAL PROPERTIES OF THE BETA PDF

The Beta pdf takes on only positive values and can be symmetrical and unsymmetrical. The function lies between the boundaries of 0 and 1. It can be scaled by a multiplication factor to represent the distribution of a random variable greater than unity. The probability $p(x)$

of the Beta pdf is defined by the following function [10]:

$$p(x) = \frac{x^{\alpha-1} \cdot (1-x)^{\beta-1}}{B(\alpha, \beta)} \text{ for } 0 \leq x \leq 1$$

$\alpha > 0$ and $\beta > 0$

$$\text{where } B(\alpha, \beta) = \int_0^1 u^{\alpha-1} \cdot (1-u)^{\beta-1} du \quad \dots (1)$$

The mean (μ) and variance (σ^2) of the random variable can be calculated from α , β and the scaling factor (b) using the following expressions:

$$\mu = b \cdot \frac{\alpha}{\alpha + \beta} \text{ and } \sigma^2 = b^2 \cdot \frac{\alpha\beta}{(\alpha + \beta)^2 (\alpha + \beta + 1)} \quad \dots (2)$$

Furthermore α and β can be calculated from the mean and variance as follows

$$\alpha = \frac{\mu(b\mu - \mu^2 - \sigma^2)}{b\sigma^2}$$

$$\text{and } \beta = \frac{(b - \mu)(b\mu - \mu^2 - \sigma^2)}{b\sigma^2} \quad \dots (3)$$

3. LOAD PROFILE SIGNAL

Any load profile can be treated as a time domain signal. In signal theory the power represented in any signal is readily calculated by the root mean square (RMS) of the sampled signal values using the following equation:

$$Y_{rms} = \sqrt{\frac{1}{N} (I_1^2 + I_2^2 + I_3^2 + I_4^2 + \dots + I_N^2)} \quad \dots (4)$$

N = The number of data points or sample size.

I = The sample data load current mean points.

The criterion for accurate representation of a signal is the sampling frequency which must comply with the minimum Nyquist theorem (sampling frequency $> 2 \times$ highest signal frequency). A load profile is merely an indication of the load central tendency over a period.

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only the poles, or consumers, or transformer or any combination of these.

4.5 Output

DataVis gives the designer the option of not exporting the data back into CART 3.0. This allows for non-destructive work on the transformer zone and the ability to use DataVis as a simple 3D World viewer. DataVis checks to see that the objects exhibiting electrical properties are placed on the ground (or above the ground in the case of conductors) and notifies the designer before exporting of data. This is the only check that DataVis performs because the designer can upon importing the file from DataVis into CART 3.0 check the connections. DataVis outputs a text file that contains the junction pole co-ordinates, feeder cable lengths and service cable lengths. CART 3.0 will read this file and update the 2D CAD drawing with any changes made.

5. TESTING AND EVALUATION

The DataVis prototype was implemented with two main objectives in mind. Firstly it was used to validate the use of the Distributed Virtual Reality Architecture or PAM and to measure its performance. Secondly, DataVis was developed as a tool to explore the value of a 3D Visualization tool as an aid in the development of electrification schemes.

To achieve the first objective, DataVis has been given a frames-per-second counter. In tests still to be conducted DataVis will be run on a standalone personal computer and then on a varying number of networked PCs with various process allocations. FLYBY will be used to determine the frames per second on each type of configuration (standalone or networked) and a comparison made.

To achieve the second objective, a panel of experienced electrification network designers will be set up and given CART 3.0 together with DataVis to use. Their reaction to this new tool will be measured by filling out a questionnaire after using the tool. Their perception of lag in using the tool on a standalone workstation and networked workstations will be measured.

In work still to be carried out the test results of DataVis in this environment will be compared with the findings of [2], who found with their particular tool that it was more useful for making changes to an existing design than for building a design from scratch.

6. ACKNOWLEDGEMENTS

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common element to the source path. If the algorithm has still not determined the total path, it then assumes that the destination position is in another branch of the transformer. It creates a new path with the source and destination path (without the transformer).

FLYBY gives an aerial flyby of the transformer zone based on the dimensions (furthest object in the x and z plane in both directions and height of the objects) of the transformer zone. MANIPULATE allows the user to alter object parameters. This is explained further in the next section.

4.4 Immersive Design

DataVis doesn't allow general manipulation of all of the parameters. This is because DataVis was designed as a specific visualization tool and not as a 3D Modeller (3DM [3] and Issac [5]). Junction poles can be selected and moved to new locations. The

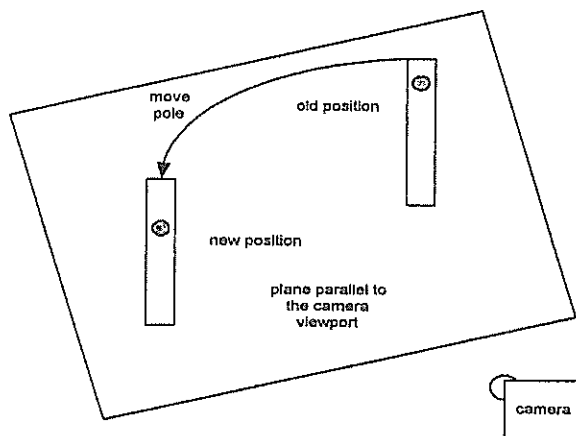


Figure 6 – Camera Viewport

closest pole to the camera viewport, as projected onto the camera viewport, will be chosen. The 2D-device space point is to be converted into a 2D-camera viewport space. Because the pole is repositioned using the mouse (which is a 2D pointing device) the only movement allowed is along a plane parallel to the camera viewport (Figure 6).

To obtain a precise position requires a steady hand and good perception (there are three dimensions to consider), therefore DataVis provides the designer with the option of moving the junction pole to a specific co-ordinate in the environment and move the junction pole onto the ground plane. DataVis will then reposition the feeder conductors attached to that pole and service conductors if the pole has consumers connected to it. This consequently alters the phase voltages and currents. The 3D bars above the junction poles consequently change in size and position themselves over the new position of the junction pole. The transformer cannot be moved

from its position because its position is selected within CART 3.0 to be as close as possible to the centroid of the transformer zone [4].

The environment allows the designer to view the expected voltages on each electrical phase at a consumer or junction pole as a percentage volt drop from the transformer. This is accomplished with 3D bars for each voltage phase where the length of the bar in the y-direction being the percentage volt drop, which are positioned above each junction pole and consumer. A red bar is for the red phase, a white bar is for the white phase and a blue bar is for the blue phase.

The CART 3.0 software [4] assumes that all electrical loads on the network are at unity power factor and thus there is no phase angle associated with the voltage. If the phase angle were something other than unity this information could be shown by rotating the bar (phase *degrees*) around the x-axis (Figure 7).

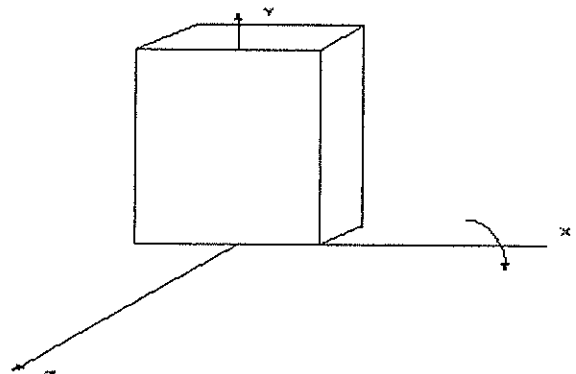


Figure 7 – Phase Rotation of Voltage Bar

DataVis allows the designer to see the voltage profile moving away from the transformer along a feeder or towards it using any of the navigation methods. In a conventional 2D representation plotting a graph showing voltage as a function of distance from the transformer would have represented this voltage data.

An important aspect that emerged from designing and implementing DataVis is that the environment should not be cluttered with useless information (objects). Users tend to navigate or steer round objects even if collision detection is not implemented. Therefore the type of objects represented in the environment was restricted to those which exhibit electrical properties. To aid in the viewing, most objects can be switched off from the rendering, thus allowing the designer to view

DataVis makes no use of the User Model object catered for in the architecture since the system presents an identical interface to anybody using the software. The Input Handler will take care of the input from the mouse and keyboard. The architecture has been expanded to provide a collision detection method in the Presentation Generator that checks whether the camera motion and the junction pole motion in the virtual environment is valid. A bounding box is placed around both objects and their intersection is checked. The Presentation Generator will also render the image onto the display. Each object process (service conductor process, junction pole process etc.) contains the following information:

- 3D model of the object
- Data for that particular object (x-y-z co-ordinates, phase voltages, type e.t.c.)
- Method for position and orientation of object

The Simulation Process handles the change of the camera's angle, position, the re-connection of the conductors to the junction poles and processes any change of viewing parameters if notified by the Input Handler.

4.2 Rendering

DataVis makes use of Criterion Software's RenderWare™ API to render the environment. The API handles the mathematics behind rotating, scaling and translating objects. The API takes care of hidden surface removal, clipping, palette management and shading. The objects in the environment are texture mapped. DataVis takes care of collision detection so that the user cannot go through cables, poles or consumers.

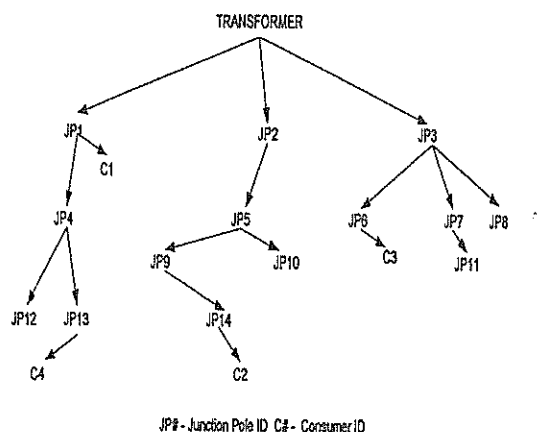


Figure 5 – "Christmas" Tree

4.3 Navigation

DataVis allows the electrical parameters in the distribution network to be visualised by allowing the designer to "move" around the circuit watching how these parameters vary spatially. The environment is navigated using the desktop mouse device. Pressing the left-mouse button and moving the mouse allows the user to pan the camera. While pressing the right-mouse button and moving the mouse tilts the camera. Positioning the mouse over a junction pole provides the user information in a status box at the bottom of the screen. The label of the pole, the number of consumers on each electrical phase "downstream" from that pole, the distance from the transformer, the x-y-z co-ordinates of the pole, and the voltages of each phase at the pole are shown in the status box. Positioning the mouse over a low voltage feeder or service conductor displays conductor-specific information in the status box.

The navigation tools and manipulation tools that are applicable to any object can be accessed from the menu or by double-clicking the right mouse button. This brings up a pop-up menu, which will position the camera at positions decided by the user. The user can drop onto the terrain and experience the feeling of walking on the ground and viewing the environment and its properties from that perspective. "Hotspots" can be defined so that a quick button press on the keyboard will move the user to that point. The designer can "fly" to a particular object, the further the object is from the current position the faster the "flying" will be.

There are three modes of navigation, namely WALK, MANIPULATE and FLYBY. WALK allows the user to transverse the environment either freely or forces a specific route by designating a start and end position. The user can instruct the software to start at a consumer or pole and move to some other pole or consumer. The software renders a path from the source (start) position to the destination (end) position

Navigation along the network is accomplished by using information the system holds about the connectivity of the various objects in the environment. It is assumed that the network is connected in a radial or Christmas tree-like structure (Figure 5). The transformer is the "angel" on top of the tree, the conductors and feeders are the branches, the junction poles are the nodes and the consumers are the leaves in the tree. The network-tracing algorithm follows a path from the source and destination positions back to the transformer. It then determines whether the destination position is in the source path. If not, it then checks to see if the destination path has any elements in the source path. If so, it appends the destination path from the

platform and because most designers of electrification networks work on Windows-based PCs.

The designer is presented with a 2D CAD drawing in CART 3.0 showing the town plan of the area to be electrified. He places symbols representing transformers, poles and junction points on the map. Lines can be drawn which represent the conductors that connect each consumer to a transformer. CART 3.0 has tools that can be invoked to allow the designer to determine the voltage at any point in the network and the current in any conductor. There are also tools to prepare detailed bills of materials and an accurate cost of the network.

DataVis allows the designer to visualize the electrical quantities within a single transformer zone. A transformer zone is an area that encloses all those consumers connected to a single transformer. A typical transformer zone consists of about 35 junction poles (and conductors to connect them to other junction poles) and 90 consumers (and conductors to connect them to junction poles).

DataVis can be invoked from within CART 3.0. A text file containing the following data is produced for the currently selected transformer zone:

- Geographic co-ordinates (x-y co-ordinates) of each consumer's connection point, each pole and the transformer;
- The connectivity between the transformer, poles and consumers;
- Details of each pole (height, label and length of incoming conductor) and the number of consumers connected to each electrical phase on the pole.
- Descriptions (type, resistance, maximum current rating, etc.) of each feeder and service conductor.

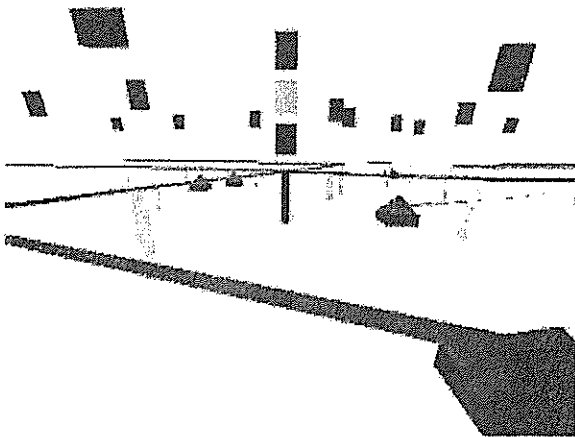


Figure 3 - Snapshot of DataVis

A Virtual Environment is produced from reading in the data from the text file as seen in Figure 3. The x-y co-ordinates are mapped onto the x-z plane in the 3D environment and the y co-ordinate of the 3D environment is derived from the heights of the poles as read from the text file. A ground plane is added to the environment as a reference plane. The transformer is positioned at the centre of the environment with all the other object's positions defined relative to the transformer's position. This allows for simpler calculations while navigating (moving from one object to another) within the virtual environment.

In setting up the virtual environment the positions and orientations of the feeders and service conductors have to be calculated. These are calculated by knowing the positions of the source pole and the pole or consumer to which the respective conductor must be attached.

4.1 Distributed Virtual Reality Architecture

The DataVis software has been implemented using the Distributed Virtual Reality Architecture or PAM as described earlier. DataVis was first developed for a standalone workstation. The architecture was then added to the DataVis software.

The DataVis program is loaded on each of the networked PCs used in the simulation. DataVis knows which computers are available for the simulation. The architecture will assign processes to various computers. The DataVis software residing on the computer running CART 3.0 will notify the other computers in the simulation about the allocation of processes. A certain PC may, for example, handle the Presentation Generator, another may handle the Input Handler and several others

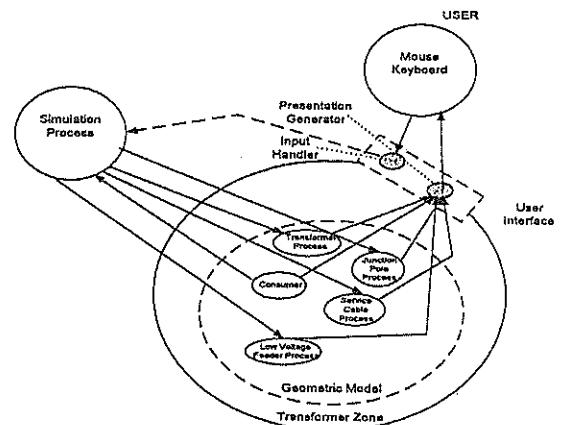


Figure 4 - Virtual Reality System Model for DataVis

may have the object processes reside on them. This allocation of processes is invisible to the user.

electrical network. Using VR, however, it would be possible to allow the designer to interact with the electrical network within a realistic 3D environment in which important parameters such as voltage and current are represented visually.

VR and other advanced MMI techniques usually require expensive high performance graphical workstations. The approach adopted in the research project presented in this paper has been to develop a "Distributed VR Software Architecture" [1] that allows computationally intensive advanced MMI techniques to be implemented on a network of standard "office" personal computers. The financial resources required to set-up a truly virtual design environment with all the head-mounted displays, gloves and other sensory-feedback devices is large. Therefore the project is concerned with using equipment found on standard personal computers such as graphical colour displays and the mouse input device.

3. DISTRIBUTED VIRTUAL REALITY ARCHITECTURE

To increase the performance of a Virtual Reality application, it can be distributed across a number of networked processors using a software architecture we have called the "Process Agent Model" (PAM) [1]. The PAM makes details of the distributed processing and network communications transparent to the high level processes and to the user.

The message passing mechanism used between remote machines is the same as that used between processes residing on the same machine so that an identical executable resides on each computer. On each computer there exists either the authentic process or its agent process. The agent process acts as representation of the authentic process. The agent intercepts messages destined for the process it represents and passes these messages through the network to the authentic remote process.

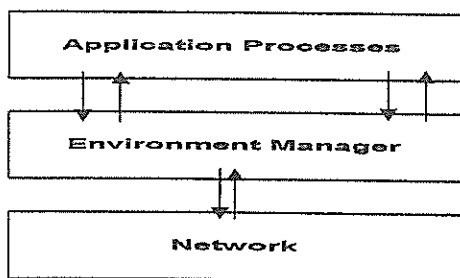


Figure 1 – Three Layered Architecture

The architecture is layered so that it hides the underlying communications from the objects above.

The Network, the Environment Manager and the Application Processes constitute the layered architecture (see Fig.1).

The Environment Manager takes care of system initialisation and process to processor allocation. It also handles the routing of messages from agents to the remote computer housing the authentic process. The network communications infrastructure makes use of raw Ethernet frames to communicate between remote process on remote computers. An intelligent dual incoming queue mechanism handles the incoming and outgoing traffic on each computer. The mechanism is intelligent in that only the latest update from each process is retained.

The interface between the user and the virtual world (Figure 2) consists of the Presentation Generator (output) and the Input Handler (input). A number of object processes together form the geometric model. The Input Handler may initiate changes to the Object Processes. The Input Handler can also provide inputs into the Simulation Process. The Geometric Model incrementally communicates net changes to the Presentation Generator, which in turn renders the virtual environment on the display. The Simulation Process handles the simulation aspect, which may receive information from the Input Handler, and from objects in the Geometric Model. The conceptual system model decouples the simulation process from the rest of the system.

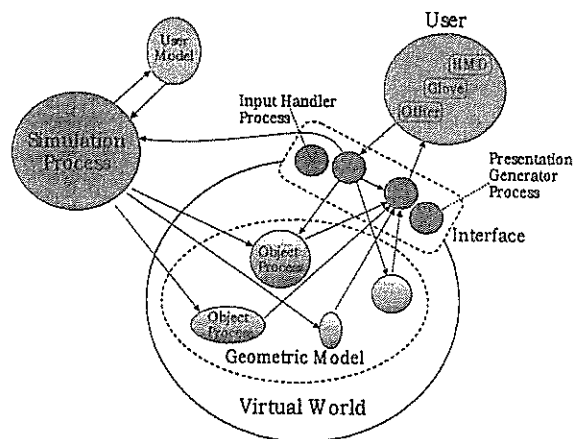


Figure 2 – Virtual Reality System Model

4. VISUALIZATION

The DataVis software interfaces with an integrated set of tools, called CART 3.0 [4] developed to assist the designer of electrification schemes. DataVis can be run on any standalone or networked computers that are capable of running Microsoft Windows™ NT or 9x. This platform was chosen because the PAM (see previous section) was designed on this

VISUALIZATION OF TRANSFORMER ZONES USING A DISTRIBUTED VIRTUAL REALITY ARCHITECTURE

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Abstract: In South Africa and other developing countries there is currently a drive to electrify homes of disadvantaged communities. To aid the electrification process requires effective software applications. The set of stand-alone tools currently used are inherently based in two dimensions and convey a limited amount of information. This paper introduces a software application, DataVis, which allows designers to visualize a three-dimensional environment from a two-dimensional CAD drawing. The application allows for the viewing and manipulating of electrical and physical properties of the objects in the environment. A distributed virtual reality software architecture is used to allow networked low-cost personal computers to create and allow the designer to interact with this environment. The concept of using data visualization in three-dimensions will be investigated for its applicability in the electrification design process.

Keywords: data visualization, virtual reality application, distributed virtual reality, electrical reticulation design tool.

1. BACKGROUND

A major electrification program is currently underway in South Africa connecting 1000 houses per day to the national electricity grid. The majority of these new connections are being made in informal settlements close to major cities and in rural areas. Designers are faced with keeping the cost of the designs of the electrification systems as low as possible while maintaining high construction standards. Computer tools obviously have a significant role to play in supporting designers in this task.

2. INTRODUCTION

The design process currently undertaken by the designers of large low-cost electrical distribution networks depends on a range of non-integrated computer tools. The town plan of the area to be electrified is prepared on a two-dimensional CAD system. The layout of the network is usually developed in pencil on hardcopies of the CAD drawing. Specialised stand-alone software is used to calculate voltage levels and currents. Once all design decisions and considerations have been finalised a draughtsman will then produce a detailed construction drawing on a CAD system.

Over the past few years emphasis has been placed on developing software tools that streamline and improve the design process with the goal of optimizing the design [4]. The integration of all the design activities within a common software

environment eliminates manual re-entry of data and other time-consuming and wasteful tasks. The designer makes use of a common software environment to quickly explore different design options and thus produce cheaper networks.

Many designers in South Africa are beginning to use an integrated software environment developed around a commercial CAD package [4]. A major limitation of this two-dimensional design tool, however, is that it is difficult for the designer to deal with the large amount of data relating to voltages, currents and costs. The move from a paper-based design methodology to an integrated CAD-based one has not had a significant impact on the visualization of data. The designer still sees voltage and current values represented as numerical text on a flat drawing of the town's layout. Using the graphical capabilities of the current generation of computers, it becomes apparent that they can be used for the visualization of data in various ways leading to a better understanding of a design.

This paper deals with an ongoing research project which aims to explore the application of advanced MMI techniques to the visualization of data within an electrical distribution design environment. Virtual Reality (VR) allows us to interact with a reality that does not exist outside the environment of the computer on which it runs. VR allows us to visualize data in ways that would be impossible within the physical world. We cannot, for example, see the variations of voltage and current in an actual

ply frequency, it is used in industry for measurement and protection. Current transformers used for measurement must be accurate under normal conditions, whereas for protection, current transformers must maintain their accuracy at the highest levels of fault or waveform distortion.

Current transformers operate similarly as other transformers with magnetic cores. Excitation of the primary winding sets up a change of flux in the core. Ignoring losses, the secondary winding sets up a change of flux, equal in magnitude but reversed in direction to oppose this change in flux.

This simplified (and ideal) description can be adapted to account for secondary effects due the materials and construction methods used. For example, lumped modeling of a lossy linear core, stray capacitance and leakage inductance can be added to this ideal model. A lumped model of a current transformer can have many components, therefore a simplified model is used for a specific application. An important aspect is to know to what accuracy and under what circumstances such a model is valid.

The model shown in Figure 2 is a lumped model of a current transformer, taking into effect certain secondary effects. The primary and secondary resistance and leakage inductance, core loss and magnetizing inductance are elements of the model. At supply frequency, a simplified model of a current transformer, neglecting C_e and C_d , can be used. At high frequencies, the magnetic coupling becomes weak, and the signal is mostly passed due to the capacitance of the winding.

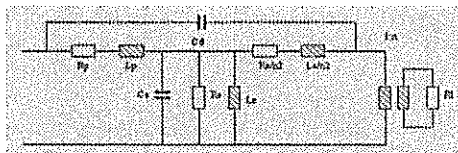


Fig. 2. Model for Current Transformer

L_m	magnetizing inductance
R_e	core loss
R_s/n^2	secondary resistance
R_p	primary winding resistance
C_e	stray capacitance
C_d	stray capacitance
L_s/n^2	secondary inductance
L_p	primary leakage inductance
R_l	load resistance
Legend for Figure 2.	

B. Rogowski Coil

The Rogowski Coil is a coil set up in a configuration to measure the current passing through the coil. A current carrying conductor sets up a flux directly proportional to the current. The output voltage of a coil is directly proportional to the differential of the flux. By integrating this voltage, the current can be obtained. One method to determine the output signal of the Rogowski coil can be calculated using Ampere's circuital law

$$\oint \mathbf{B} d\mathbf{S} = \mu \mathbf{I}$$

where μ is the permeability of the medium. For a toroid with n turns, pitch Δs , and circular winding area A (radius ρ), contour length ℓ , the emf induced due to field variation $\frac{d\mathbf{B}}{dt}$ is

$$v(t) = \frac{\mu n A}{\ell} \frac{di(t)}{dt} = M \frac{di(t)}{dt}$$

By this method, however, the flux induced by an external or non-centric current carrying conductor cannot be evaluated. For a single turn in a time-varying magnetic field, the emf generated is given by

$$emf = \frac{-d\Phi(t)}{dt} = \oint \mathbf{E} \cdot d\mathbf{L} = \frac{-d}{dt} \int \mathbf{B} \cdot d\mathbf{S}$$

Let the Rogowski coil be a toroid with M winding (N turns each) around the z -axis. Figure 3 shows such a configuration. The toroid has a radius R to the center of the area S_m (perpendicular to ϕ) enclosed by a winding.

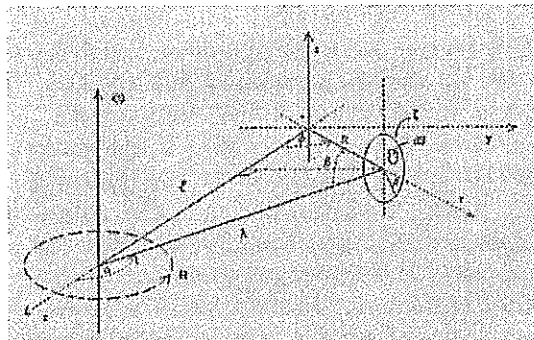


Fig. 3. Geometrical Configuration of Coil

In Figure 3 define dS_m to be the m -th surface element equiangular around the z -axis. Let ℓ be the distance from a conductor to the origin. For a long straight conductor the magnetic field intensity due to the current at a distance is given as

$$\mathbf{H} = \frac{i(t)}{2\pi\lambda} \phi$$

where $B = \mu H$ in free space. It can be shown that

$$emf = \frac{di(t)}{dt} \left(\frac{-\mu_0}{2\pi} \right) \int_{-\zeta}^{+\zeta} \int_{R-\rho}^{R+\rho} \frac{\cos \beta}{\lambda} (dr dz)$$

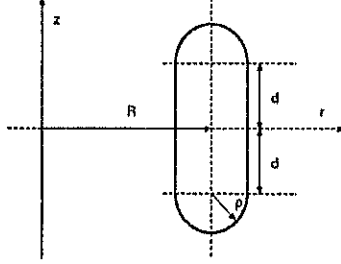


Fig. 4. Rogowski Coil Winding Shape

For the shape as shown in Figure 4, elongated in the z -axis direction with a distance of $2d$, the limits of integration (ζ) becomes $\zeta = d + \sqrt{\rho^2 - (r^2 - R^2)}$. The induced voltage V is given by

$$V = N \frac{\mu_0}{\pi} \sum_{m=0}^{M-1} \int_{R-\rho}^{R+\rho} \frac{\zeta(r - \ell \cos \phi_m)}{r^2 - (2\ell \cos \phi_m)r + \ell^2} dr \frac{di(t)}{dt}$$

where $\phi_m = \frac{2\pi m}{M}$. This type of integral does not have an analytical solution [1], hence numerical methods have to be used. The parameters for the coil is 5 equiangular spaced coils with 25 turns each, $\rho = 6.5mm$, $d = 62.5mm$ and $R = 30mm$. The output voltage is $emf = 1.4878 * 10^{-6} \frac{di}{dt}$.

IV. MEASURING SENSORS PERFORMANCE

A. Current Transformer

In this experiment the time behaviour of the CT when saturated is shown. A $50Hz \times 25A$ signal was applied to a $2.5VA$ CT with a current ratio of $15 : 1$. A load of 4.4Ω was applied to this CT causing the CT to exceed the rated VA . Figure 5 shows the input and output current of the current transformer under this condition. This figure indicates clearly that the CT is driven into saturation. This clearly constitutes an inappropriate use of the current transformer.

The frequency response of the CT has been evaluated with the aid of measuring system as described above. A distorted current was injected into the primary of the CT. Simultaneous measurements were made of the input and output currents. Low inductance current shunts with a resistances of $0.97m\Omega$ and $5.00m\Omega$ was used for the primary and secondary currents respectively. A 16 bit converter with a sampling rate of $1250Hz$ per

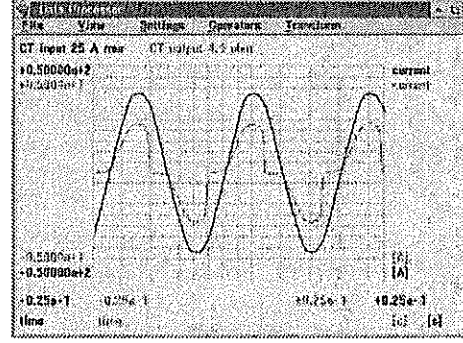


Fig. 5. Input and output current of CT under saturated conditions.

channel was used for data capturing. Figure 6 shows the input and output currents of the current transformer.

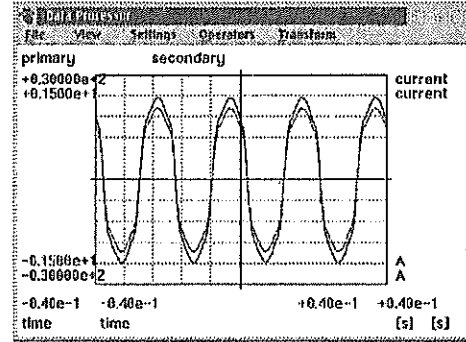


Fig. 6. Input and output current of CT.

The current $i(t)$ with waveform defined as

$$i(t) = \sum_{n=1}^5 A_n \left(1 + \sin(n\gamma_{2n-1}t - \frac{\pi}{2}) \right) (\sin n\omega_{2n-1})$$

with values $A_1 = 1$, $A_2 = 0.1$, $A_3 = 0.1$, $A_4 = 0.07$ and $A_5 = 0.05$ was applied. The value for γ is 0.625π and the value for ω is $2\pi 50$.

The current $i(t)$ was injected into the primary of the CT. Figure 7 shows the output plotted against the input for the test wave when applied to the current transformer. The primary and secondary current of the CT was measured using low inductance current shunts with $0.97m\Omega$ and $5.00m\Omega$ resistances respectively. A straight line will reflect that the output is a true replica of the input.

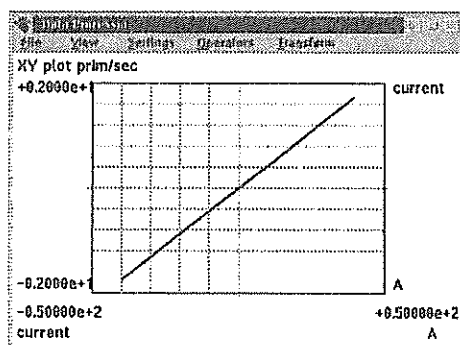


Fig. 7. Output versus input of the Current Transformer.

f [Hz]	I_{in} [A]	I_{out} [A]
50	1.0000	1.000
150	0.1323	0.1325
250	0.0838	0.0840
350	0.0383	0.0386
Current transformer normalized input and output frequency components.		

B. Rogowski Coil

Figure 8 illustrates the relationship between the input current and output voltage of the Rogowski coil. Figure 9 illustrates the frequency components of the Rogowski coil voltage output.

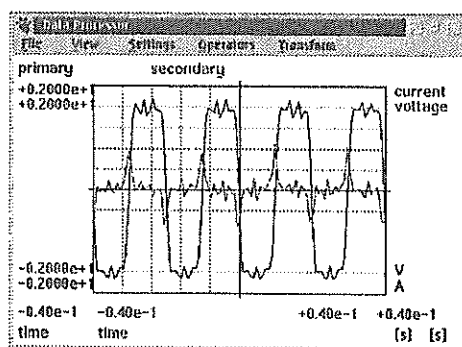


Fig. 8. Rogowski coil input current and output voltage.

f [Hz]	I_{in}	V_{out}	$\int V_{out}$
50	1.000	1.000	1.000
150	0.2719	0.7999	0.2666
250	0.1208	0.5823	0.1164
350	0.0917	0.640	0.0914
Rogowski Coil normalized input and output frequency components.			

V. CONCLUSIONS

It can be concluded that the current transformer that has been tested, can be an accurate current sensor for power quality assess-

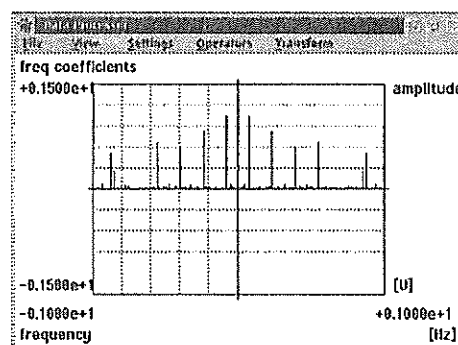


Fig. 9. Frequency components of Rogowski coil output.

ment if the CT is operated with a low secondary impedance and the VA rating are not exceeded. The output signal from the Rogowski coil is very small resulting in a poor signal to noise ratio. The signal to noise ratio is much better at higher frequencies. Due to the expected harmonic profile typically of a power system, the Rogowski coil accuracy steadily improves at higher frequencies. The current sensors should be characterized for all the possible operating conditions, so that power quality assessment can be done accurately.

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PROCEDURE FOLLOWED FOR EVALUATING THE VOLTAGE DIP (SAG) PERFORMANCE OF VSI-TYPE VSIDS TESTED AT THE UNIVERSITY OF THE WITWATERSRAND VOLTAGE DIP TESTING FACILITY

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Abstract: This paper describes the process of voltage dip testing at the University of the Witwatersrand voltage dip testing facility. An introduction to the voltage dip test facility is firstly given. Types of voltage dips tested and other factors, which could influence the drive voltage dip performance, are discussed. The procedure followed when performing voltage dip testing on drives is described and an overview of the method used to present the voltage dip test results is given.

Keywords: power quality, voltage sags, variable speed drives, power system reliability

1. INTRODUCTION

Voltage dips are a common cause of plant disruption and are the result of a fault at one point on the electricity supply network causing the voltage at another point to vary. During voltage dips industrial equipment can function abnormally [1], but as the voltage dip performance of equipment is usually unknown the effect of voltage dips on equipment cannot be quantified. A voltage dip test facility has been developed at the University of the Witwatersrand in conjunction with Eskom for determining the voltage dip performance of equipment. Variable speed drives (VSDs) are used in many industrial applications but, being electronically based, can be susceptible to power disturbances. Voltage dip testing, at a test facility such as the one at Wits University, can be performed on these drives to determine the drive response during voltage dip conditions. Results from these tests can be used by utility customers to determine the effect of voltage dips on their plant and by drive manufacturers to improve the voltage dip performance of their equipment.

2. OVERVIEW OF THE VOLTAGE DIP TEST FACILITY

The test setup used for testing the ASD is shown in Fig. 1 and its development is discussed in [2]. A PC is used to synthesise a complex voltage dip by adjusting magnitude and phase parameters in the dip synthesis software module for each of the three phases independently. The synthesised waveform is loaded into the voltage dip testing software module running on the testing PC. Under user control firstly a user-defined pre-dip waveform is applied to the equipment and when the equipment is operating normally the synthesised voltage dip is applied. Equipment measurements are fed into the PC and during the

voltage dip the testing software logs these measurements to disk where they are stored for later analysis. For the VSD testing performed, input voltage, input current, DC bus voltage and motor speed were recorded. Loading of the VSD was accomplished by coupling the induction motor to a DC generator supplying a resistive load. The test procedure is repeated to investigate the effect of different voltage dip types and drive operating conditions.

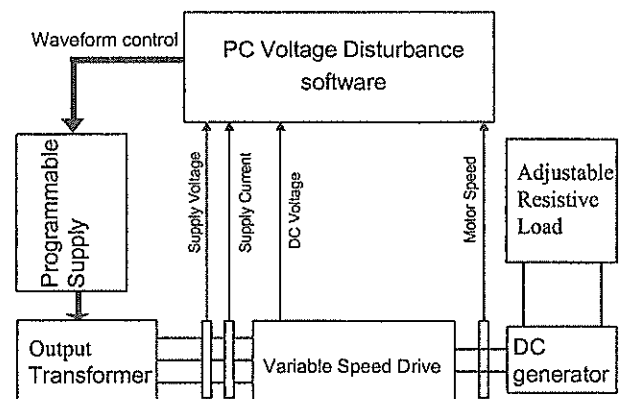


Fig. 1: Block diagram of the test facility when testing an adjustable speed drive

3. DESCRIPTION OF THE DRIVE TO BE TESTED

A block diagram of a voltage source inverter (VSI) type drive is shown in Fig. 2. The single-phase or three-phase mains supply to the drive is rectified and filtered with a DC bus capacitor. Inductors may be present on the line-side (line reactors) to reduce the current harmonics generated by the drive rectifier. A DC link inductor is sometimes placed on the output of the rectifier supplying the capacitor. The DC bus is

connected to an inverter output stage which supplies the induction motor. At the present time this is usually an IGBT-based inverter output stage. Although the basic topologies are similar in different drives, drive size and drive manufacturer can influence the exact topology. For example, a power factor corrected input stage may be added to improve the power factor and reduce line harmonics typically associated with a diode rectifier.

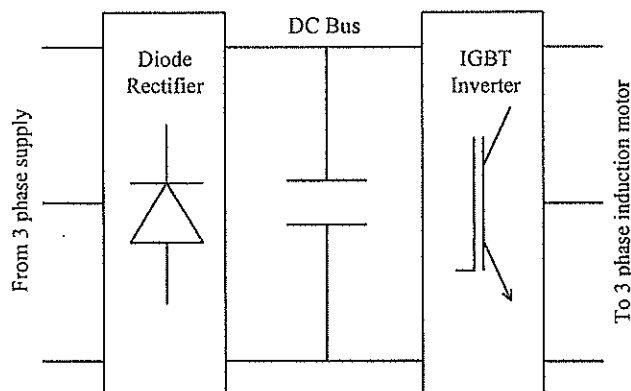


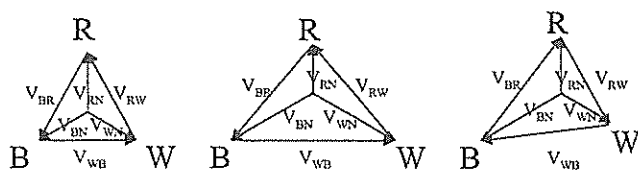
Fig 2: Block diagram of a typical VSI PWM VSD.

Although the topology of these types of drives are similar, voltage dip performance is dependent to a large degree on component sizes used by the manufacturer, control strategy used by the manufacturer and the undervoltage sensing strategy implemented. For these reasons drives of similar specification may have vastly different dip performance behaviour.

3. VOLTAGE DIP TESTING ON THE VSD

3.1. Types of voltage dip

The voltage dips applied to the drives tested are shown in Fig. 3. These voltage dips are characterised according to the number of phases affected during the voltage dip: single-phase affected, two-phases affected and all three-phases affected. For the testing performed only magnitude and duration of the voltage dips was adjusted.



(a) 3-phase dip (b) 1-phase dip (SLGF) (c) 2-phase dip

Fig. 3: Types of voltage dips tested

For the single phase and two-phase voltage dips, there are three possible voltage dip combinations depending on the individual phase(s) affected. Some drives may behave differently depending on the phases(s) affected during the dip. In order to characterise the dip performance of the VSD this implies that the drive

must be tested with all seven possible combinations of voltage dip type.

3.2. Pre-dip voltage and drive loading effect on dip performance

When testing a particular VSD at the University of the Witwatersrand voltage dip testing facility, other factors, which could influence the dip performance of the drive, are investigated. In particular the loading of the drive and the pre-dip voltage are varied. The tests undertaken used the following values:

- Drive loading, 25% and 75% of rated ASD load.
- Pre-dip voltage, 0.95 p.u., 1.0 p.u. (nominal) and 1.05 p.u.

4. Testing Procedure

For each of the types of voltage dips applied to the drive, the equipment is tested to find the dip magnitude at which the drive trips for different dip durations. Typically the dip duration is varied between 1 cycle (20 milli-seconds at 50Hz) and 3 seconds. Fig. 4 shows a flow chart of the procedure followed when performing voltage dip testing.

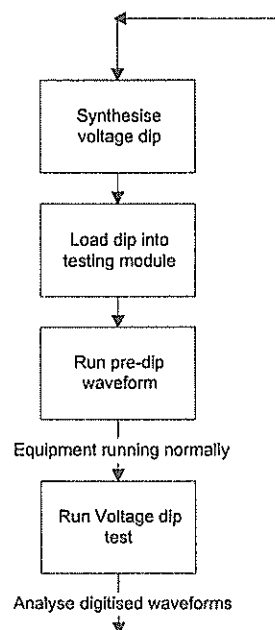


Fig. 4: Procedure followed when performing test

For a particular dip type, load value and pre-dip voltage a series of tests is undertaken for different dip durations. For each dip duration value the dip voltage is varied until the dip magnitude when the drive trips is established. The drive tripping may be an actual drive undervoltage trip or may be a change in drive operating parameters (e.g. motor speed) that is deemed unacceptable. Drive measurements logged during the voltage dip test can be analysed following the test to determine if a particular parameter varied out of preset limits.

5. PRESENTATION OF TESTING RESULTS

If all the different permutations of voltage dip type, drive loading and pre-dip voltage are tested, this implies 42 different types of test will need to be undertaken. For each of these tests the dip duration is varied from 20 milli-seconds to 3 seconds and the dip magnitude is varied to establish the trip threshold for that particular test and dip duration.

For testing performed at the University of the Witwatersrand, the results from each test type are plotted on a dip depth Vs duration graph. To reduce the number of graphs, different drive loading test results are plotted on the same graph. For the testing outlined above typical testing results in a set of 21 dip-depth versus dip duration curves for each of the voltage dips tested. Fig. 5, Fig. 6 and Fig. 7 show results for each of the three types of voltage dip tested. [3] discusses comprehensive voltage dip testing performed on a particular VSD.

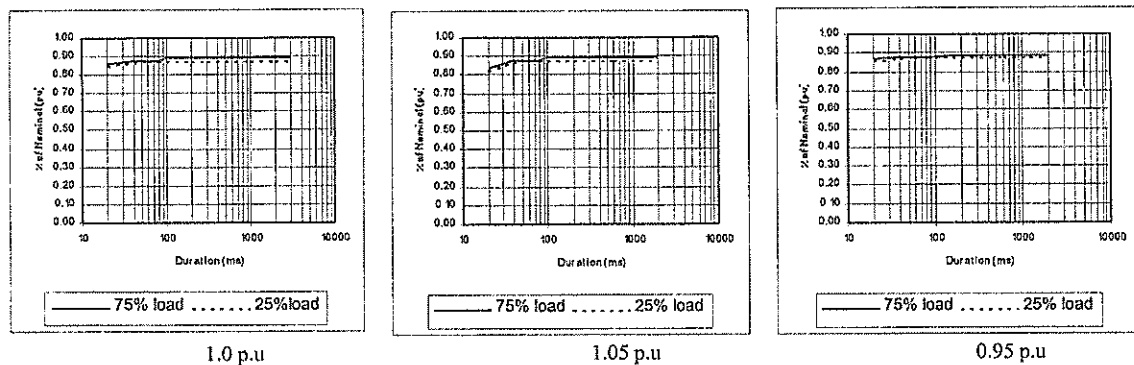


Fig. 5: Three-phase voltage dip ASD ride-through performance

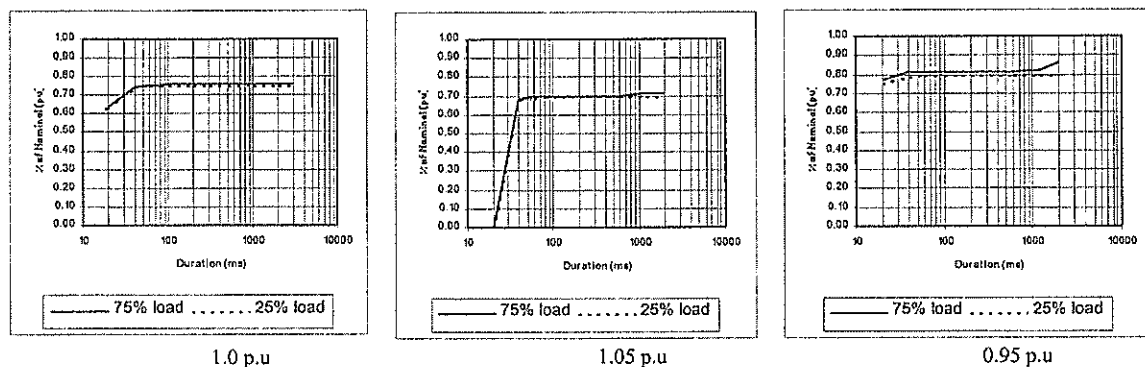


Fig. 6: Single-phase (Blue) voltage dip ASD ride-through performance

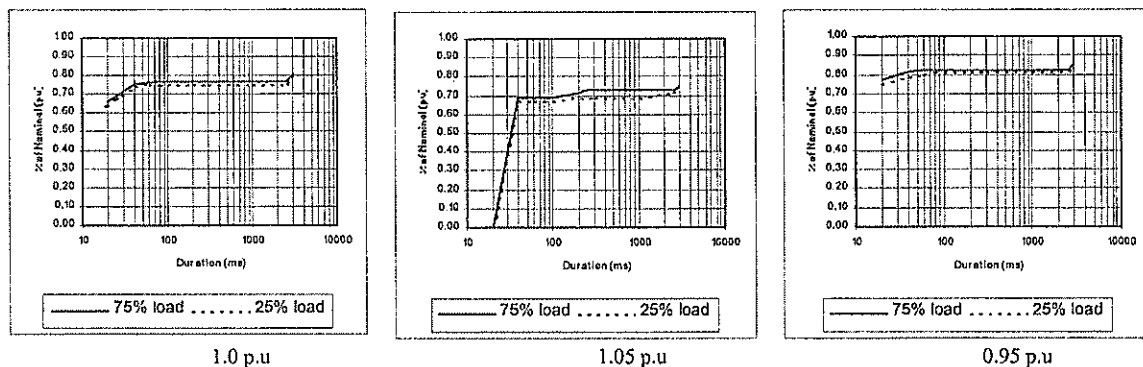


Fig. 7: Two-phase (Red-blue) voltage dip ASD ride-through performance

6. CONCLUSION

This paper described the test procedure followed when performing voltage dip tests on variable speed drives at the University of the Witwatersrand dip testing facility. Results from these tests allow customers to specify certain dip performance criteria which can then be confirmed by testing. From the test results, customers can also make informed decisions on dip mitigation strategies. Knowledge of the drive dip behaviour allows customers to avoid drives with sensitivities that coincide with predicted or expected voltage dips.

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A NOVEL METHOD FOR THE GENERATION OF VFTS AT HIGH PRESSURE IN A GIS TEST VESSEL.

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ABSTRACT

A new method of generating VFTs at high gas pressure is presented. A capacitor bank is discharged through the primary of a gas insulated transformer. This generates a switching impulse at the secondary terminals of the transformer with a rise time of approximately 1 ms. A disconnector switch is used as a sharpening gap to produce the initial step to excite the VFT in the experimental test setup. The amplitude of the VFTs produced are dependent on the gas pressure in the disconnector chamber, the disconnector contact spacing and the voltage on the capacitor bank. This new system allows for easy installation and small size. The VFTs produced have amplitudes up to 1.55 MV. V-t curves for a 180/110 mm co-axial gap at 3 and 4 bar absolute are presented. The shape of these curves conformed to those of past work at lower voltages.

Key Words: GIS, VFTs, V-t Curves, SF₆, high gas pressure

1. INTRODUCTION

SF₆ technology enjoys a highly respected service record for voltages up to 800 kV. However, problems may arise from the operation of the disconnector switches [1]. These switches are used to isolate off-load parts of the station, for maintenance purposes. Very Fast Transients (VFTs) will be generated by the operation of a disconnector switch. The peak magnitude of the VFT can reach two times the operating voltage of the system. VFTs have been known to cause failures in systems below the lightning withstand level of the system.

initial step with a rise time of less than 5 ns. When flash over occurs across the disconnector gap, travelling waves are transmitted from the switch in both directions along the Gas Insulated Switchgear (GIS). At any impedance discontinuity (a spacer or joint between adjacent sections), part of the travelling wave is reflected. These reflections sum to form the complex VFT. The characteristics of the particular VFT are unique to the particular test set-up used and will be different for different test set-up configurations.

2. EXPERIMENTAL METHODS USED FOR GENERATING VFTS

2.1. The Previous Method

Previously in our laboratory, VFTs were generated by energizing one section of the test set-up using a gas insulated transformer and then closing the disconnector switch onto the un-energized section of the test set-up. The transformer was supplied from an auto-transformer with parallel and series SCRs. The drive to the series SCRs would be switched off when a breakdown in the test vessel was detected to prevent supplying short circuit current to the transformer. Gating signals are applied to the parallel SCRs when a breakdown in the test vessel is detected to short circuit the transformer primary. Due to the large quantity of EM noise produced at higher voltages when there is a breakdown in the test vessel, the control panel was proving unreliable. As a result of this, VFT V-t curves could only be completed up to voltages of approximately 800 kV which implied a pressure limit of 2,5 bar absolute [2, 3].

2.2. New method

The new method of VFT generation relies on producing a slowly rising switching impulse out of the secondary of the transformer and letting the disconnector gap flash over to produce the VFT. The slowly rising switching impulse out of the secondary of the transformer is

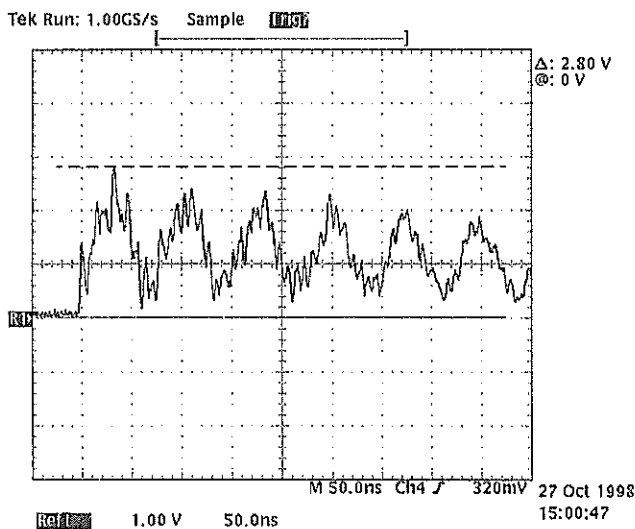


Figure 1. A typical positive polarity VFT generated in the experimental set-up of peak magnitude 658 kV. Horizontal 50 ns/div. Vertical 235 kV/div.

The VFT in Figure 1 rises to its first peak in approximately 35 ns and has an oscillatory top with a frequency of approximately 13,6 MHz. The rise time of first initial step is dependent on the gas pressure in the disconnector. The magnitude of this first initial step is dependent on the ratio of the surge impedance on either end of the disconnector switch contacts. This VFT has an

generated by discharging a capacitor bank into the primary of the transformer. The capacitor bank was charged up to approximately 1800 V. A high voltage contactor is used to switch the charged capacitor bank onto the primary windings of the test transformer. This causes a steep-rising voltage across the primary windings, which is stepped up to the terminals of the secondary winding. The inductance of the transformer slows this injected voltage to a switching impulse with a rise time of approximately 1 ms at the secondary terminals.

The disconnector switch is left open at a predetermined contact spacing. The disconnector contacts will flash over producing the VFT. The peak amplitude of the VFT is proportional to the voltage across the disconnector prior to breakdown across its contacts. A higher disconnector chamber gas pressure would therefore cause a larger VFT for the same disconnector contact spacing. Figure 2 shows the VFTs produced with the disconnector chamber pressure at 3 and 7 bar respectively, with the same disconnector gap spacing.

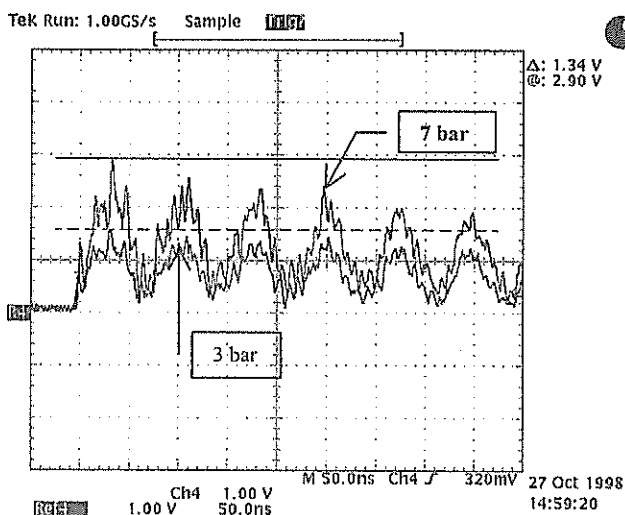


Figure 2. VFTs produced with the disconnector chamber pressure at 3 and 7 bar respectively, with the same disconnector gap spacing. Horizontal 50 ns/div. Vertical 235 kV/div.

This new method of VFT generation has the following advantages:

1. There is no need for the bulky control desk and its associated series and parallel SCR combinations to protect the transformer.

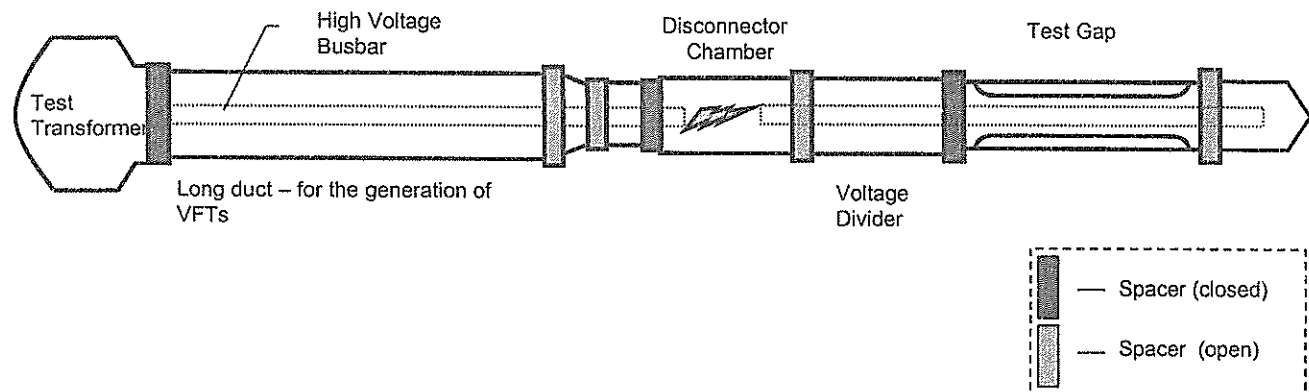


Figure 3. This figure shows the test pole used to generate the VFTs.

2. The power requirements are quite low. Only the capacitors need to be charged.
3. The capacitor bank and associated circuitry is far more compact than the control desk.
4. The charging voltage can remain constant and the disconnector alone can be used to control the magnitude of the VFTs.
5. The new method allows for VFTs to be produced with peak magnitudes up to 1500 kV. VFTs of this magnitude make it possible to construct V-t curves at pressures up to 5 bar in the case of the test duct. This method of VFT generation is seen as a tool with which research can progress at higher pressures, which would be relevant to industrial applications.

Reversing the polarity of the diode in the rectifying circuit changes the polarity of the generated VFTs. VFTs of the chosen polarity are guaranteed.

3. EXPERIMENTAL SET-UP

The test pole used is the same as that used by Reynders and Traynor [4]. It consists of industrial grade GIS sections. The test gap is a 180/110 mm coaxial geometry. The test pole is shown in Figure 3. A capacitive voltage divider in an adjacent disconnector chamber is used to measure the generated VFT voltage waveform. The voltage divider has a division ratio of 235 000 : 1. The divider and oscilloscope combination have a bandwidth of 500 MHz.

A 142 μ F capacitor bank with a rating of 2500 V was used. The capacitor bank was charged up to approximately 1800 V. A high voltage contactor is used to switch the charged capacitor bank onto the primary windings of the test transformer

4. Generated VFT V-t curves

The method used to plot the V-t curves is that proposed by Reynders [4]. A VFT V-t curve plots points that relate the highest voltage before or at breakdown to the time to breakdown. A line is drawn through the lowest points to generate the VFT V-t curve. This line denotes the voltage level, below which the GIS will not fail. Information regarding the characteristics of the VFT used to generate the V-t curve must be supplied along with the curves as

the shape of the VFT is dependent on the geometry of the installation.

4.1. Correction factors

A 10 % correction factor was added to the measured values to offset the effects of remote measurement [5].

4.2. Test results

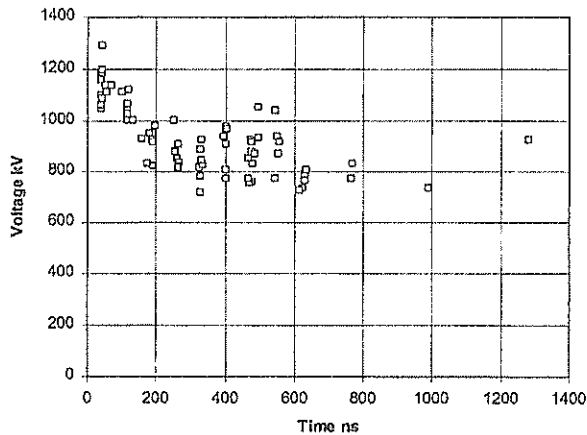


Figure 4. The negative polarity VFT V-t curves for the 180/110 mm coaxial gap at 3 bar absolute.

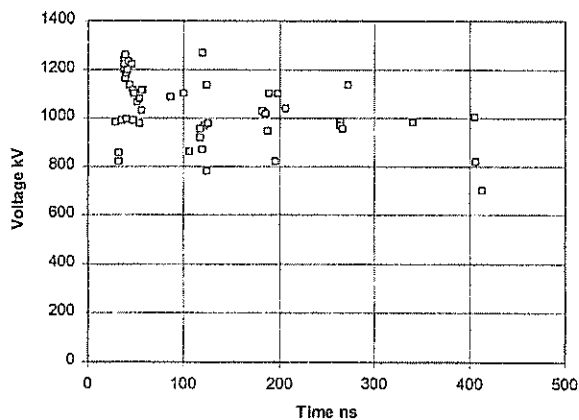


Figure 5. The positive polarity VFT V-t curves for the 180/110 mm coaxial gap at 3 bar absolute.

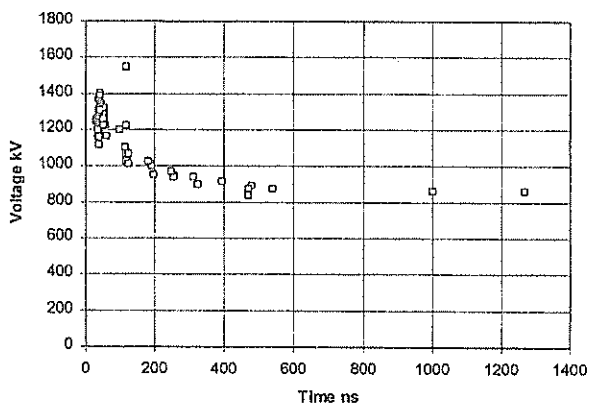


Figure 6. The negative polarity VFT V-t curves for the 180/110 mm coaxial gap at 4 bar absolute.

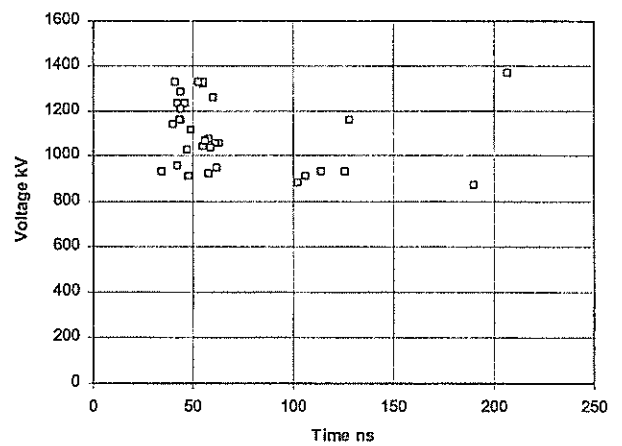


Figure 7. The positive polarity VFT V-t curves for the 180/110 mm coaxial gap at 4 bar absolute.

V-t curves for positive and negative VFTs were constructed and can be seen in Figure 4, Figure 5, Figure 6 and Figure 7 for 3 and 4 bar absolute. The predicted streamer breakdown values are 735.4 kV and 976.6 kV for 3 and 4 bar absolute respectively.

5. DISCUSSION OF RESULTS

The flatness of the positive VFT V-t curves with respect to the negative curves is evident. The reason for this is that breakdown initiated by positive VFTs has a relatively short formative time lag. This leads to the majority of breakdowns occurring on the first peak of the VFT, shortly after the critical voltage for breakdown has been reached.

Breakdown caused by negative voltages has a larger formative time lag. This could be due to the thicker discharge channel that is present in the case of negative breakdown. It may also be due to the nature of how the electrons used in the breakdown are “freed up”. For negative voltages, the electrons are pulled out of the conductor, whereas for positive voltages they are the result of the ionization of the gas molecules. This implies that the work function of the conductor is higher than the ionization energy of the gas molecules. As a result of the larger formative time lag, low probability breakdowns occur up to some hundred nanoseconds delayed on the subsequent oscillation, which leads to the distinctly up turning trend of the corresponding V-t curve.

Both positive and negative curves smooth off below the calculated value for breakdown voltage. This is due to the field enhancement caused by surface roughness. The oscillating tail of the VFT is said to lower the leader inception voltage, this would lead to lower than critical voltage breakdowns for longer breakdown times as leader breakdown characteristics are slower than streamer breakdown characteristics.

It proved exceedingly difficult to get positive VFT breakdown times greater than 400 ns at 3 bar and 200 ns at 4 bar. There was no problem getting longer times for negative polarity VFTs. This is once again due to the

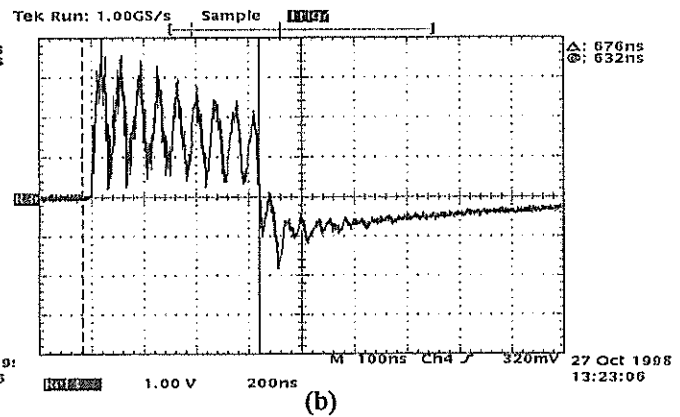
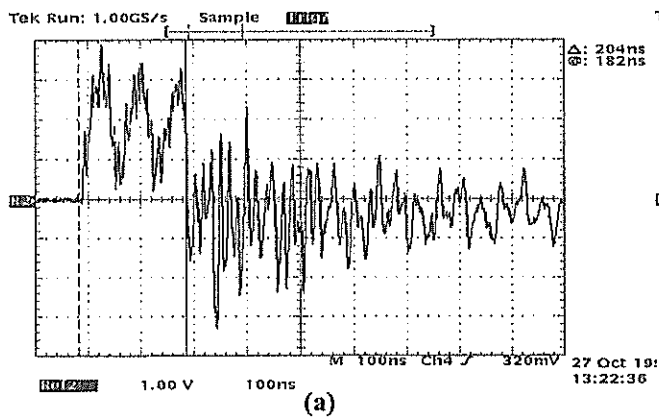


Figure 8. a) Breakdown in the test gap. Horizontal 100 ns/div. Vertical 235 kV/div. b) "Problem" flashover. Horizontal 200 ns/div. Vertical 235 kV/div.

longer formative time lag (and therefore "delayed" breakdown times) present in breakdown under negative voltages.

5.1. Problems with the Test Procedure

During the tests conducted for positive VFTs at 3 bar, breakdown occurred in the test pole, but not in the test duct. The VFT was seen to be significantly different. After breakdown in the test duct, the measured voltage oscillated around the zero level. With this new phenomenon, the voltage drops below zero for positive VFTs and decays back to zero after more than a microsecond. The two traces can be seen in Figure 8.

The "problem" breakdown only occurred if the test duct hadn't broken down after about 500 ns. This points to a slower breakdown mechanism causing the fault. If in fact the oscillating tail of the VFT reduces the leader inception voltage [6], it is possible that leader type breakdown is causing the faults. This could be occurring in two places in the duct:

1. Between the disconnecter contact and ground. This is, however unlikely as the electric field between disconnecter contacts is uniform, which would promote streamer-type breakdown as opposed to leader.
2. Over a spacer. This is the more likely cause of the problem breakdown. Carbon deposits were found on one of the spacers.

6. CONCLUSIONS

A new method of producing VFTs is presented which allows for the generation of high amplitude VFTs. The peak magnitudes of the VFTs are dependent on the voltage across the disconnecter contacts prior to breakdown across the gap. As a result of this, the magnitude of the VFTs generated can be controlled by altering the contact spacing as well as by altering the disconnecter chamber pressure.

The equipment needed for the generation of the VFTs is

simple and can be configured in a small area in a short time.

The VFTs produced have amplitudes up to 1550 kV, which makes the construction of V-t curves at high pressures are possible.

V-t curves for a 180/110-mm co-axial gap arrangement are constructed for SF₆ pressures of 3 and 4 bar absolute. The effect of formative time lag on the shape of these curves is clearly evident.

Breakdowns outside the test gap can be clearly distinguished from those in the test gap.

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DEVELOPMENT OF A 250 kW DYNAMIC POWER QUALITY COMPENSATOR

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Abstract: This paper describes an industrialized 250 kVA Dynamic Power Quality Compensator (DPQC). The DPQC is a shunt power quality device functioning as an Active Power Filter and Dip Compensator. The Dynamic Power Quality Compensator is a multi-functional device performing power quality conditioning under normal supply conditions while being able to support the load for the duration of short dips in the utility supply using local energy storage. The multi-functional application of this power quality device results in increased device utilization and return on investment. This forms a cost-effective method of power quality compensation for sensitive industrial loads. The shunt version of the DPQC is discussed in this paper.

Keywords: Power Quality Compensator, Uninterruptible Power Supply

1. INTRODUCTION

An increasing number of industrial loads, such as paper and steel mills, are sensitive to the quality of their electric supply, while utilities are required to support growing loads using existing infrastructure. The sensitivity of various loads to the quality of their supply and associated financial losses have lead to a growing awareness of the quality of the electrical supply and methods of improving the quality of the supply [1], [2], [3].

A typical application of the Dynamic Power Quality Compensator (DPQC) for the improvement of the quality of the electrical supply to a sensitive section of a client plant is shown in Fig. 1.

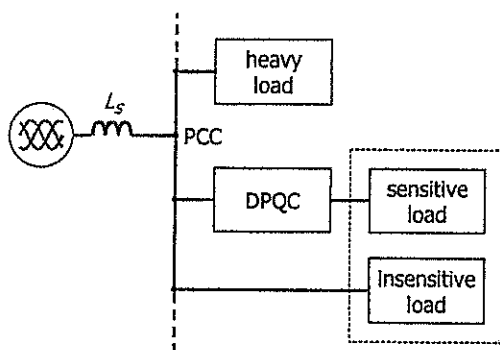


Fig. 1: A typical application of the DPQC

The power quality sensitive section of the plant, shown in Fig. 1, may be adversely affected by fluctuations in the voltage at the point of common coupling (PCC) as result of activities at heavy plants or line-faults on the same feeder. The majority of dips and interruptions are of short duration (<20 seconds) and have a negligible effect on loads with long time-constants such as heating equipment, but is detrimental to sensitive loads such as variable speed machine drives.

It is cost-effective to support only the sensitive section of the load with the DPQC as shown in Fig. 1. The DPQC can be configured as a shunt or series power quality compensating device. The shunt version of the DPQC is discussed in this paper.

2. SHUNT DPQC HARDWARE AND TOPOLOGY

The DPQC is a multi-functional device consisting of an IGBT-based inverter with the associated filter elements, a thyristor-based isolation switch, energy storage and switch-gear controller by a high-performance DSP-FPGA-based controller [7]. A simplified representation of the shunt version of DPQC is shown in Fig. 2.

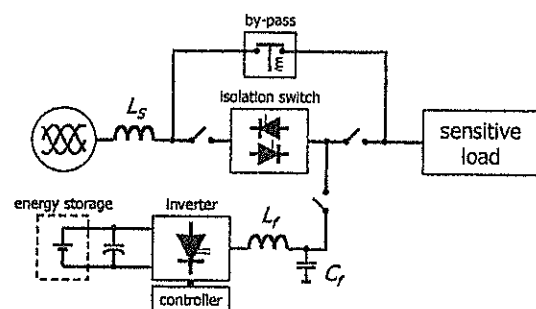


Fig. 2: Topology of the Shunt version of the DPQC

2.1. Power Stage

The DPQC power stage consists of a 3-phase inverter with intelligent IGBT power modules as switching elements. The design of the power stage allows the addition of a 4th power module enabling the DPQC to support unbalanced loads [4].

The power stage is rated for operation at 400 Vrms with a continuous load current of 400 Arms from operating dc-bus voltage of 800 V.

2.2. Thyristor Isolation Switch

A thyristor-based isolation switch is used, with the by-pass contactor in the open state, to isolate the DPQC and the sensitive load from the utility in the event of a dip in the utility supply.

During normal supply conditions, with the DPQC in the active mode, the thyristor isolation switch remains closed while the by-pass contactor is in the open state. In the event of a dip in the utility supply the thyristor isolation switch can be opened within milliseconds, isolating the DPQC and sensitive load from the utility for operation in the UPS mode.

2.3. Energy Storage

The DPQC can be directly interfaced to battery, flywheel and super capacitor energy storage devices. In this paper an 800 V, 100 Ah battery energy storage system was used. The batteries are charged under normal supply conditions and are used as an active power source during the Dip Compensation mode of operation.

High quality sealed lead-acid batteries are used, reducing maintenance and enabling operation in hazardous environments. The energy storage is designed to support a load of 250 kW for up to 30 seconds in the UPS mode.

2.4. Switch-gear

The DPQC is equipped with input and output circuit breakers (MCBs), a by-pass contactor, an inverter output and a storage contactor.

The input and output MCBs are used for the protection of the DPQC/load in the event of faults on the internal/external bus bars. The MCBs may trip as result of an over-current condition, manual switching by the operator or excitation of the shunt-trip coil by the controller. If an MCB has entered the tripped state, the DPQC requires intervention from the operator to recover.

The by-pass contactor may be used to de-activate the DPQC while the load is still supported from the utility supply. It should be noted that the DPQC is not able to operate with the by-pass contactor in the closed state and will not respond in the event of a dip in the utility supply. The DPQC enters the by-passed state when the device is deactivated by the operator or as result of an error condition. The operator may force the DPQC to enter the by-passed mode using a manual by-pass switch on the front panel.

The inverter output and storage contactors are controlled by the DPQC software. The contactors are

used during soft starting and to isolate the power stage and energy storage when the DPQC is deactivated by the operator, or as result of an error condition.

2.5. Controller

A high-performance DSP-FPGA-based controller was developed for the control of the DPQC. A TMS320C31 floating point DSP, capable of up to 50 million floating-point operations per second is used for the implementation of control algorithms, mode selection and protection.

A fixed-point DSP (TMS320C50) controller was developed to function as a human interface. The human interface consists of a graphic liquid-crystal display (LCD) and a mimic-display indicating error conditions, the current state of the switch-gear, operational parameters, remote access etc.

The DPQC enclosure containing the IGBT power stage, filter elements, thyristor isolation switch, switch-gear and the controllers are shown in Fig. 3(a). The IGBT power stage and controller are shown in Fig. 3(b).

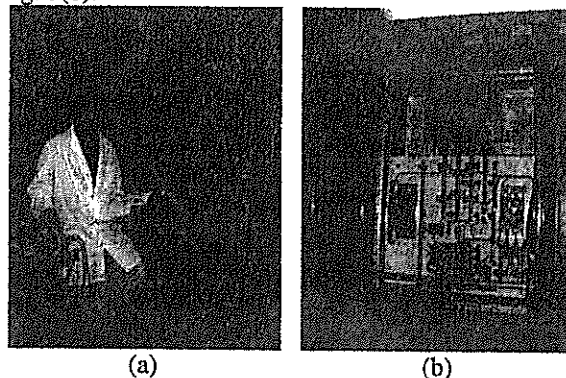


Fig. 3: (a) The DPQC enclosure, (b) Power Stage and switch-gear

3. MODES OF OPERATION

Two main modes of operation are defined for the DPQC, the UPS or dip compensation mode and the Shunt Power Quality Compensation mode. The primary function of the DPQC is the protection of the sensitive loads against detrimental fluctuations in the supply voltage by entering the UPS mode when required. The DPQC performs the secondary function of shunt power quality conditioning and re-charging of the energy storage under normal supply conditions.

3.1. Dip Compensating

The primary function of the DPQC is to protect the sensitive load against dips and short interruptions in the electrical supply of up to 30 seconds. In the event of a dip in the supply voltage, the DPQC and the sensitive load are isolated from the utility supply by opening the thyristor isolator. The sensitive load is

now supported by the DPQC operating in the UPS mode, as shown in Fig. 4.

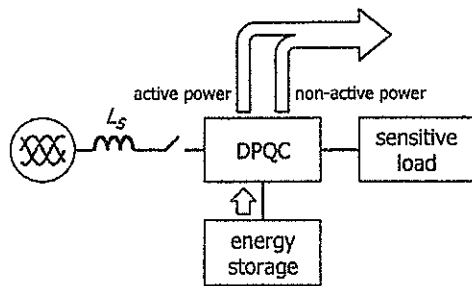


Fig. 4: UPS/Dip mode of operation

As shown in Fig. 4 the active and the non-active load power is supplied by the DPQC in this mode and no power flows from the utility supply. In this mode active energy is supplied by the energy storage. The DPQC remains in the UPS mode until a valid utility supply is detected or the energy storage is depleted.

When a valid utility supply is detected, the DPQC synchronizes the load voltage to the supply voltage before closing the thyristor isolator and entering the Shunt mode of operation.

3.2. Shunt Power Quality Compensation Mode

Under normal supply conditions when UPS or Dip Compensation functionality is not required, the DPQC performs the secondary function of shunt power quality conditioning and charging of the energy storage as shown in Fig. 5.

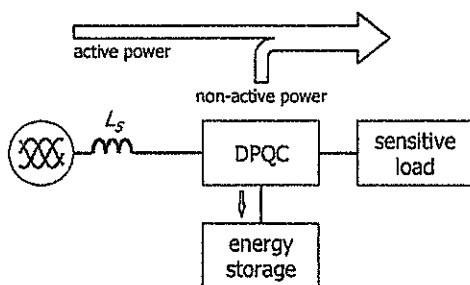


Fig. 5: Shunt Power Quality mode of operation

As shown in Fig. 5 the DPQC supplies only the non-active component (reactive and distortion power) of the load power while the active component of the load power is supplied by the utility. Additional active power is consumed by the DPQC to cover internal losses and for charging of the energy storage. The DPQC remains in this mode of operation until the occurrence of a dip in the utility supply, when the UPS mode is entered.

4. CONTROL

The DPQC is controlled by a high performance DSP-FPGA based controller equipped with digital I/O

system and a measurement system sampling numerous current, voltage and temperature measurements. The DSP and the FPGA are used in collaboration enabling various options for protection and task specialization.

High performance space vector orientated control algorithms are implemented including a predictive current regulator [7], symmetrical current limiting in the UPS mode of operation and advanced techniques for synchronization and dip detection. The APF control algorithm is an adapted version of the Instantaneous Reactive Power Theory (IRPT) suggested by Akagi [8].

4.1. Dip Detection

The strategy implemented for the detection of supply dips is illustrated in Fig. 6. The 3-phase supply voltages (V_{sa} , V_{sb} , V_{sc}) are represented by a space vector in the d-q plane (V_{sd} , V_{sq}) and the magnitude of this vector ($|V_s|$) is monitored by the DPQC controller.

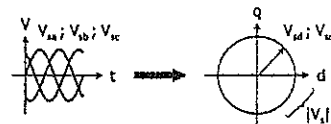


Fig. 6: Dip Detection Strategy.

Variations in the magnitude of this vector are not dependent on the phase of the vector, rendering this method insensitive to a phase shift in the 3-phase supply voltage. The magnitude of the supply vector is monitored using a digital low-pass filter, allowing an adjustable sensitivity/response time for dip detection.

4.2. Synchronization

The strategy implemented for the synchronization to the supply voltage (V_s) is illustrated in Fig. 7. The 3-phase supply (V_{sa} , V_{sb} , V_{sc}) and load (V_{la} , V_{lb} , V_{lc}) voltages used to determine two unit vectors in the d-q plane (U_{sd} , U_{sq} and U_{ld} , U_{lq}).

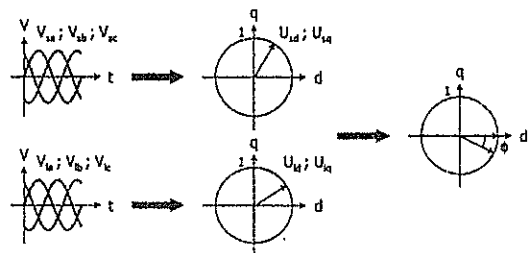


Fig. 7: Synchronization Strategy.

The DPQC controller determines and corrects the phase error using the supply and load unity vectors. This method of synchronization results in continuous tracking of the phase of the supply voltage since it does not rely on zero crossings of the supply voltages and exhibits acceptable rejection of voltage waveform distortion/noise.

5. EXPERIMENTAL RESULTS

5.1. Dip Compensating

The performance of the DPQC in the UPS mode of operation was evaluated by connecting a 240 kW load. The load consisted of a 3-phase resistive load and a 6-pulse rectifier load, each consuming about 50% of the load power.

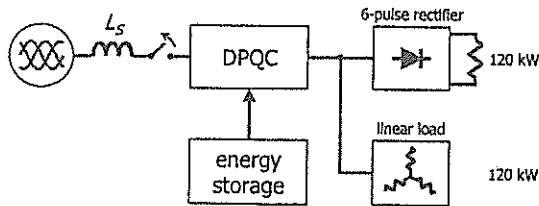


Fig. 8: Practical set-up for 240 kW Dip Compensation Test.

A dip in the supply voltage was simulated by opening a circuit breaker connecting the DPQC, and the load, to the utility supply. The phase voltage of the 3-phase utility supply, the corresponding line current and the load phase voltage were recorded. The measured results are presented in Fig. 9, where the circuit breaker was opened at $t \approx 0$.

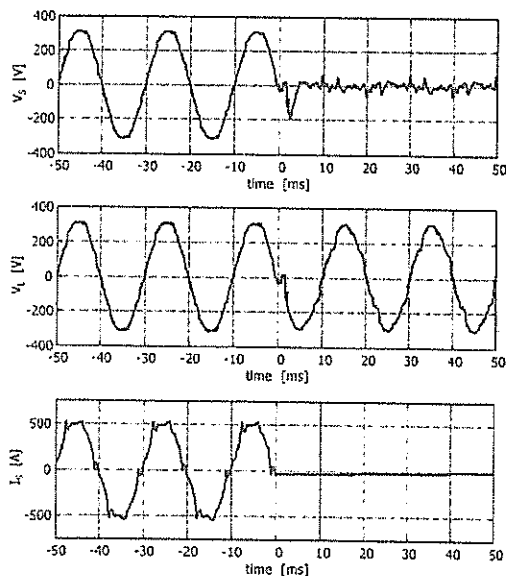


Fig. 9: UPS mode – Supply voltage, load voltage and supply current during a dip in the utility supply

The DPQC responded within approximately 4 ms, opening the thyristor isolator and entering the UPS mode of operation. The supply voltage collapsed after $t = 0$, and the supply current decayed to zero, while the voltage at the load was sustained by the DPQC operating in the UPS mode. While in the UPS mode of operation the, DPQC supplied the full load power by drawing active power from the local energy storage (see Fig. 4).

The resynchronization and the transition from the UPS to the Shunt mode of operation was evaluated by emulating a short dip using a circuit breaker. The DPQC entered the UPS mode of operation, sustaining the load for the duration of the dip. The DPQC remains in the UPS mode until a valid utility supply is detected. After synchronizing the load and utility voltages, the thyristor isolator is closed and the Shunt mode of operation is entered. The measure phase voltage at the load and the supply line current is presented in Fig. 10, where the thyristor isolator was closed at $t \approx 0$.

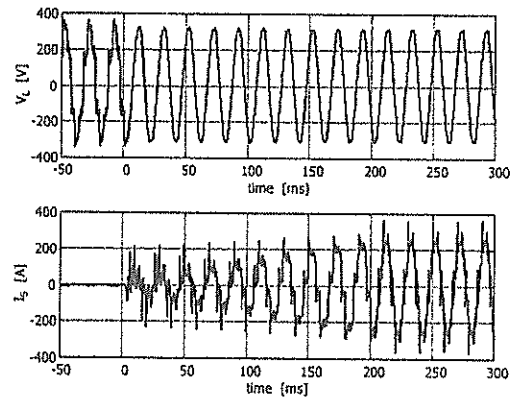


Fig. 10: UPS mode – Load voltage and supply current during re-synchronization

The fundamental component of the supply current increased to the steady-state value over a period of about 10 supply cycles, reducing transients in the supply voltages and currents during re-connection to the utility supply.

5.2. Shunt Mode

The performance of the DPQC in the Shunt mode of operation was evaluated by connecting a 6-pulse diode rectifier as a load. The load current and the compensated supply current were recorded and are presented in Fig. 11.

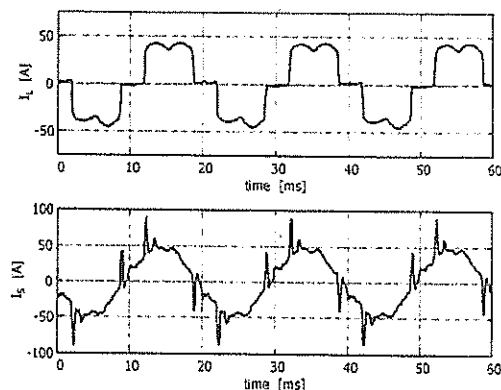


Fig. 11: Shunt mode – Load current and compensated supply current during normal supply conditions

The measured load current shown in Fig. 11 conforms to the current waveform of a typical inductively loaded 6-pulse diode rectifier. The compensated supply current contains spikes corresponding to the occurrence of very high slopes in the load current. This is the result of delays in the control algorithms, data acquisition system and the limited dynamic capability of the inverter.

The 6-pulse diode rectifier load operated at a high displacement power factor and no significant phase shift is observed between the load current and the compensated supply current presented in Fig. 11.

6. CONCLUSIONS

It has been experimentally verified that the DPQC, a shunt power quality device, can be successfully configured to operate as a multi-functional device. Increasing device utilization and thereby increasing the return on investment, is a cost-effective power quality solution, beneficial to both the client and the utility [2], [8].

The client benefits from the use of the DPQC since it results in:

- A higher quality and reliability of the electrical supply – lower losses from interrupted manufacturing processes and damaged products as resulting from equipment malfunction as result of dips in the electrical supply.
- Operation at an improved displacement power factor – an improved displacement power factor lead to supply currents of lower magnitude resulting in lower cost.

The utility benefits from the use of the DPQC since it results in:

- Improved customer satisfaction – a higher quality and more reliable electrical supply is presented to the client, especially in regions experiencing problems with frequent dips of short duration.
- Reduced non-active load current components – the load current has an improved displacement power factor and a reduced harmonic contents which is favorable attributed for loads on a network with limited infrastructure.

The DPQC may be interfaced to various types of energy storage devices such as batteries, flywheels and super capacitors and may be configured for operation as a shunt or series power quality compensator.

7. ACKNOWLEDGEMENTS

The authors wish to acknowledge ESKOM and AMS Power Electronic Converters for support on the industrialization of the DPQC.

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HARMONIC GENERATION BY STATIC VAR COMPENSATORS ON DISTRIBUTION NETWORKS.

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Abstract: A SVC is the least costly dynamic reactive power compensator. Harmonic voltage generation is the main disadvantage of the SVC. This paper describes the harmonics generated by a typical SVC as measured on a scale model. The specific sources of the harmonics are investigated and solutions are proposed. Its industrial suitability is also discussed.

Keywords: SVC, reactive power compensation, harmonic generation.

1. INTRODUCTION

The Static VAR Compensator (SVC) is a dynamic reactive power compensator. Reactive power compensation is required for three reasons:

1. Dynamic voltage regulation over the power system's equivalent impedance,
2. Reactive currents drawn necessitates equipment with higher current ratings and
3. Dynamic compensation of voltage flicker.

As stated in NRS 048, the onus of ensuring quality of supply is not only on the electricity supplier, but also on the consumer. The SVC is the most cost effective dynamic reactive power compensator and is designed for rapidly varying reactive power. The SVC may therefore be an excellent solution for consumers with rapidly varying loads and power factor problems.

A scale model [2], [3], was used to evaluate the SVC. Some attention is given to its ability to dynamically compensate for a poor power factor, but the focus of this paper falls on the current harmonics generated by the SVC.

If the line current drawn is non-sinusoidal, the voltage drop over the power system's equivalent impedance is non-sinusoidal. This will result in harmonic voltage components at the point of common coupling. Modern industrial equipment are increasingly sensitive to these harmonics.

Due to a mainly inductive load (e.g. line impedances, transformers and rotating machines), the power system's reactive power is usually lagging. Thyristor controlled reactors (TCRs) provide very rapid VAR compensation but need parallel capacitors to provide leading VARs [7], [5], as shown in Figure 1.

The current through the inductors are controlled by the firing angle of the thyristors. This changes their effective reactance and the reactive power which the

SVC delivers. The current through the thyristors flow only for a part of the voltage cycle, generating current harmonics. The reactor current is given by [7]:

$$I_{L1} = \frac{V_{line}}{\pi \omega L} (2\pi - 2\alpha + \sin 2\alpha) \quad (1)$$

and the effective inductance is given by [7]:

$$L_{eff} = \frac{V_{line}}{\pi I_{L1}} \quad (2)$$

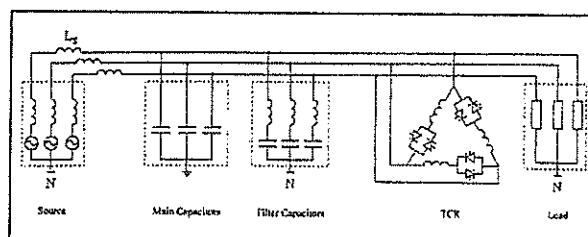


Figure 1 Thyristor controlled inductor SVC [2].

Connecting the reactors in delta, the odd triplen harmonics are cancelled. Since the phase currents are balanced, no even harmonics are generated. This leaves the fifth and seventh harmonics as the largest components. They are usually damped with tuned filters.

2. POWER QUALITY MEASUREMENTS AT PALABORA MINING COMPANY (PMC)

Reactive power and voltage harmonic were investigated at PMC, a copper mine in the Northern province. Currently their reactive power and harmonic compensation are satisfactory but additional compensation will be needed in the near future due to extensive load changes. Cyclo converter driven mine winders will be installed. These loads draw rapidly varying reactive power and cause severe voltage harmonics [7]. The relatively low cost and dynamic reactive power compensation of the SVC is ideally suited for this application but the harmonics are a problem. A STATCOM costs three times more but also provides active harmonic filtering [6].

3. HARMONIC EVALUATION OF THE SVC SCALE MODEL

The scale model is a 10.4 kVA version of the 200 MVA Impala SVC in Natal. The capacitor and inductor values of the scale model are given in Table 1. The NRS 048 standards are used as a reference for the harmonics in this investigation.

Table 1 Component values of scale model SVC.

	Main	TCIs	Harmonic filters		
			3rd	5th	7th
Capacitors	10 nF		35 uF	30 uF	30 uF
Inductors		60 mH	32 mH	13 mH	7 mH

Interharmonics [8] do not exist in periodic waveforms. They are the result of aperiodic signals which are usually only spectral noise in power systems [4]. Interharmonics are therefore disregarded in this investigation.

Although the SVC's usefulness is its ability to compensate for rapidly varying loads, only steady state experiments are conducted. The reason for this seemingly incomplete evaluation is: the SVC does not compensate for the harmonics generated by the load; instead, it is expected to actually aggravate the harmonic situation. More attention is therefore given to the steady state harmonic generation of the SVC rather than its dynamic capabilities in this paper. The dynamic reactive power compensation and accompanied harmonic generation will be investigated in a future study.

The load used for the SVC scale model, is a three phase 2.4 kW induction machine. This is less than 25% of the SVC's rating of 10.4 kVA, but a very good indication of the SVC's capabilities were still obtained. A DC machine is mounted on the same shaft and is used to load the induction machine. The power developed by the DC machine is dumped in a resistor bank.

3.1 SVC not operating

Figure 2 shows the line voltage and load current for the red phase with the SVC not in operation. The phase difference between the line voltage and load current is 2 ms, giving a power factor of:

$$\cos \phi = \cos \left(\frac{2}{20} \times 360^\circ \right) = 0.809 \quad (3)$$

The current harmonics are shown in Figure 3. The Fourier transform generates side lobes and a Blackman window is used to minimise the fundamental component's side lobe interference [9].

The third harmonic indicates saturation in the magnetising current of the induction machine [1, p. 96]: the machine is rated for 380 V but driven at 400 V under full load. Eleventh and thirteenth

components, which are typical slot harmonics for an induction machine [1], are also present. The load current are balanced and therefore no even harmonics are present. The apparent second harmonic is due to the wide main lobe of the fundamental component (Blackman window).

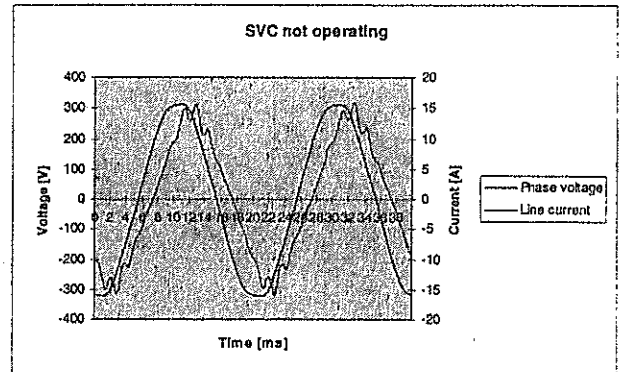


Figure 2 Line current and bus voltage with the SVC not operating.

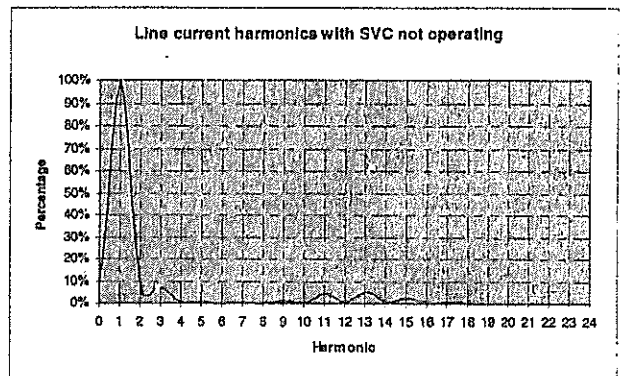


Figure 3 Harmonic content of line current with the SVC not operating.

These load conditions are kept constant throughout the experiments and serve as a reference to which the SVC's performance can be compared.

3.2 All capacitors switched in

The next step is to switch the SVC in with all its capacitors and a firing angle of 180° . This prevents large inrush currents in the inductors. This condition will deliver maximum leading reactive power. The largest harmonic currents will also be present at this condition.

Figure 4 shows the phase voltage, the line current and the inductor current of the SVC for the red phase. The power factor is unity with the firing angle at 152° .

The harmonic results in Figure 5 show a large fifth component (27% of the fundamental) and an increased third component (10%). An 8% seventh harmonic, 7% eleventh, 10% thirteenth and 2% fifteenth and 2% twenty first harmonics are also present.

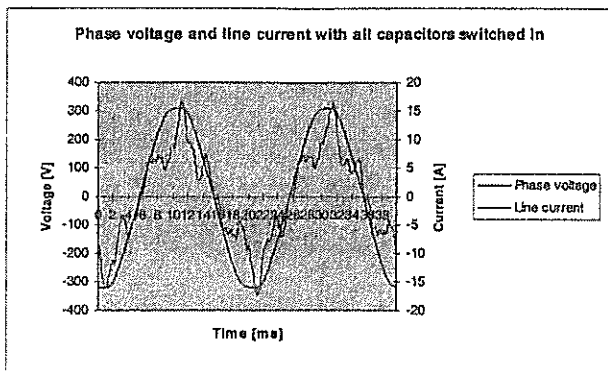


Figure 4 Line current and bus voltage with all the capacitors switched in.

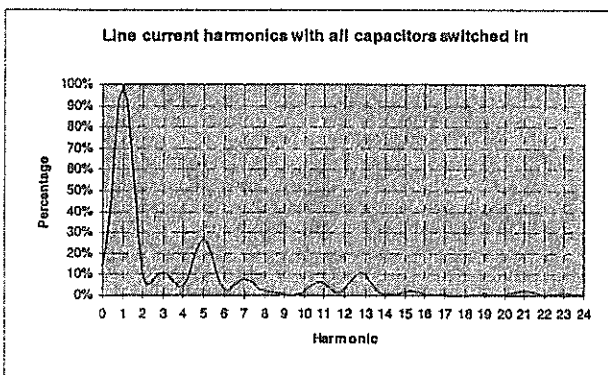


Figure 5 Harmonic content of line current with all the capacitors switched in.

The inductors are connected in delta. No triplen harmonics should therefore exit the delta connection. In Figure 6 the circulating third harmonic is clearly visible in the inductor currents. The only source of the increased third and twenty first line harmonics is therefore the capacitor banks. The eleventh and fifteenth harmonics are from the induction machine, as shown in Figure 3. The increase in the thirteenth harmonic is attributed to either the capacitor banks, the inductors, or both. The third, fifth and seventh harmonics should be filtered by the tuned filters. This discrepancy is investigated later in this paper.

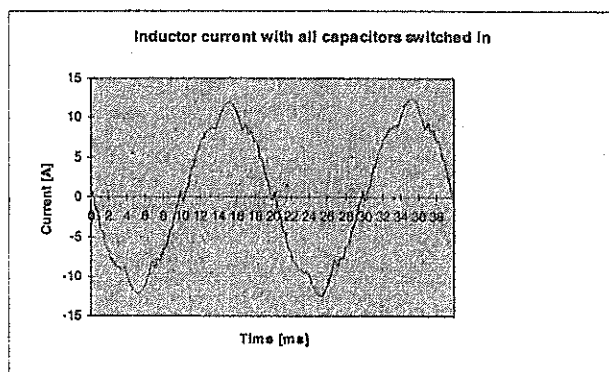


Figure 6 Inductor current with all the capacitors switched in.

3.3 Filter capacitors switched in

The main capacitors, which account for half of the SVC's leading reactive power, is subsequently switched out while the filter capacitors (the other half of the leading reactive power) remain switched in. Once again the firing angle is adjusted until the power factor is unity, as shown in Figure 7. As can be expected for a smaller capacitance, the firing angle is larger (169°) which results in a smaller inductor current, as shown in Figure 8.

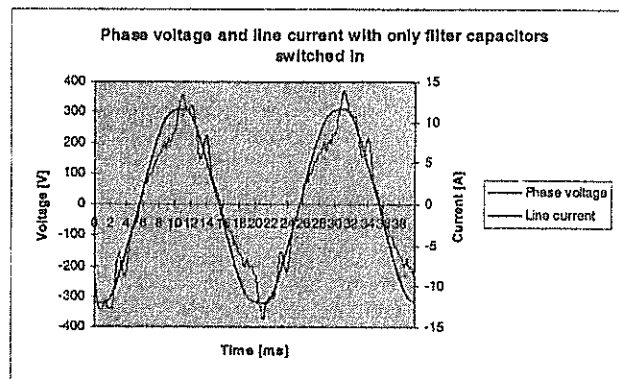


Figure 7 Line current and bus voltage with only the filter capacitors switched in.

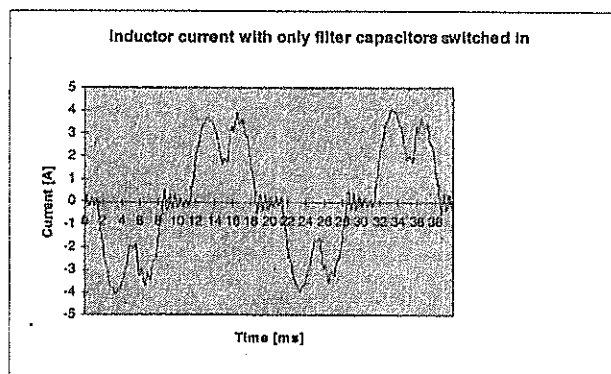


Figure 8 Inductor current with only filter capacitors switched in.

The inductor current contains a large fifth harmonic and rings when switching off. The ringing is the result of energy cycling between the effective inductance of the TCRs and filter capacitors. This resonant circuit's peak is at 1.65 kHz (thirty third harmonic) for this specific firing angle of 169° . The ringing is small and a triplen, circulating only in the delta connection. It is therefore ignored. Caution should be exercised at other firing angles as the resonance might get larger.

The harmonic analysis of the line current is shown in Figure 9. The third harmonic increased to 11% while the fifth and thirteenth harmonics dropped to 8%. The seventh harmonic remained the same. As expected, the eleventh and fifteenth harmonics from the induction machine remained the same. The conclusions are reserved until the next case is evaluated.

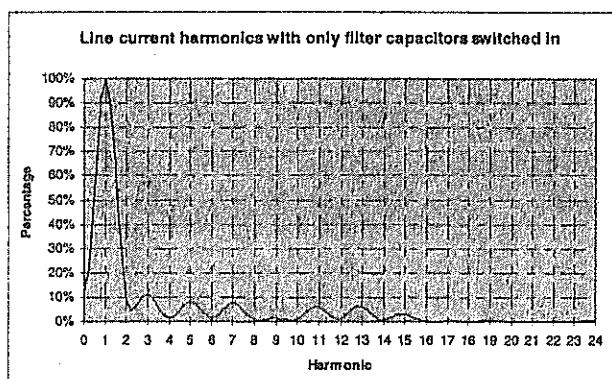


Figure 9 Harmonic content of line current with only the filter capacitors switched in.

3.4 Main capacitors switched in

Next the filter capacitors are switched out and the main capacitors are switched in. The line current and phase voltage waveforms are shown in Figure 10.

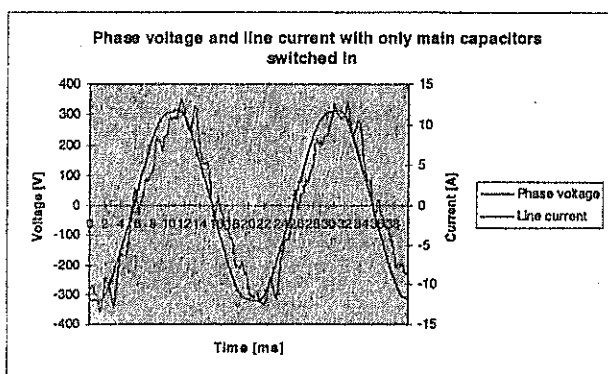


Figure 10 Line current and bus voltage with only the main capacitors switched in.

It is expected that for unity power factor, the firing angle and inductor current should be the same as for the previous case, since the effective capacitances are the same. This is confirmed by the inductor current shown in Figure 11 which compares well with the inductor current in Figure 8. The same fifth harmonic and 1.65 kHz ringing are present.

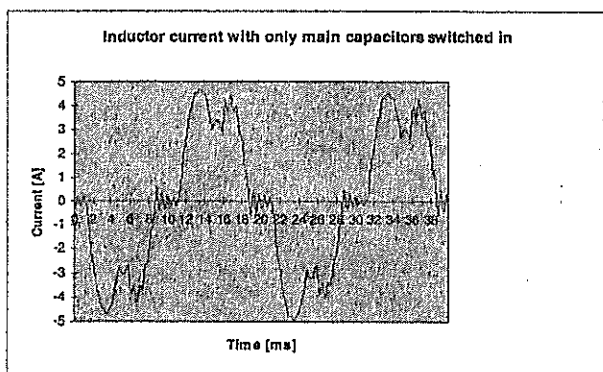


Figure 11 Inductor current with only the main capacitors switched in.

When the line current harmonics in Figure 12 are examined, a lower third and fifth harmonic are found while the seventh harmonic remains unchanged from the previous case. Once again the eleventh and fifteenth harmonics of the induction machine remain unchanged. The thirteenth harmonic has increased to 10%. The twenty first harmonic returned and it can be concluded that it is generated by the main capacitors.

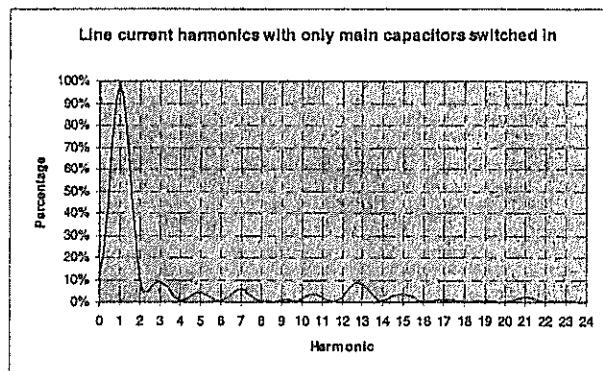


Figure 12 Harmonic content of line current with only the main capacitors switched in.

Comparing the results for the thirteenth harmonic, it is evident that the main capacitors increase this harmonic component. From the inductor current graphs and line current analysis for the last two cases, it is clear that the large fifth harmonic in Figure 5 must be attributed to the TCRs.

The fact that the filters do not sink the third, fifth and seventh harmonics are clearly a sign of them being detuned. LC filters are very sensitive to component values and can enhance harmonics rather than damping them if detuned far enough. This is the case with the third and fifth harmonics as they are lower with only the main capacitors switched in, than with only the filter capacitors switched in. The constant seventh harmonic also indicates this filter to be detuned, but not as severe as the other two cases.

4. HYBRID ACTIVE FILTER FOR SVC

A hybrid active filter, as shown in Figure 13, can be implemented in the SVC. The active filter's power rating can be small because the reactive power is still supplied by the passive components [6], [10]. The active filter dynamically tunes the passive harmonic filters.

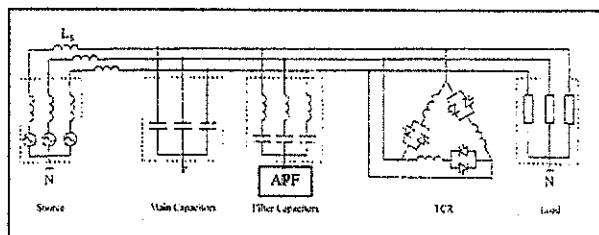


Figure 13 Hybrid active power filter for SVC.

5. CONCLUSION TO SVC INVESTIGATION

The commissioning and evaluation of the SVC are concluded by listing the main outcomes of the investigation.

- The SVC can accurately compensate for the reactive power of rapidly varying loads.
- Significant harmonics generated by the SVC are: third, fifth and seventh harmonics.
- The harmonic filters can easily detune and impair the harmonic situation.
- Filters are necessary but it is essential for them to be tuned accurately.
- Hybrid filters for SVC's can be used to dynamically tune the passive filters.
- The harmonic generation by the SVC impede its suitability in applications with harmonic problems, e.g. PMC's future loads.
- As an alternative, the STATCOM compensates both for varying reactive power and harmonics but costs about three times more.

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Experimental Validation of Loss Calculations for a Planar Inductor

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Abstract- This paper covers an experimental validation of the FEM calculations of losses in a planar inductor. The planar structure is approximated in different ways with two-dimensional FEM software to perform the loss calculations. Thermally controlled experiments are performed at different frequencies in the operating circuit to validate the theoretical FEM loss calculations for the particular operating conditions.

Keywords- Conductor losses, core losses, loss modelling, planar inductors

I. INTRODUCTION AND PROBLEM DEFINITION

The improved characteristics of modern switching devices, is significantly influencing the demand for packaging of the components ie. smaller volume, higher efficiency, better cooling. These requirements in turn influence the packaging requirements such as the electrothermal behaviour of magnetic components. Thermal behaviour of reactive magnetic components and the losses in these components have therefore experienced a significant increase in importance in the design of power electronic converters over the past few years.

The number of conference and journal papers, [1], [2], [3], [4], [5], [6], [7], [8], [9], [10], [11] investigating the modelling of losses in magnetic components used in switched mode applications, is a clear indication of this trend. The true test for acceptance of any of the modelling techniques should be thermal measurements to validate calculated loss values. A survey of a large number of papers indicates that such measurements are seldomly performed. This is ascribed to the experimental difficulties involved in accurate thermal characterization.

In this paper losses are calculated by FEM procedures and the calculated losses are verified by means of in-situ thermal measurements performed with the planar inductor in an operational inverter circuit.

II. CALCULATION PROCEDURE

The physical structure of the planar inductor investigated here is shown in Figure 2. Low profile E and I cores are used and the windings are formed on a double sided PCB.

Current commercial FEM software provide facilities to perform analyses of the complicated planar geometry and the relevant high frequency loss effects which are of interest here. Segmenting the

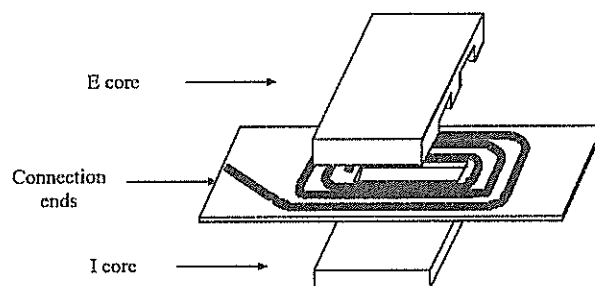


Fig. 1. Planar inductor with the E and I cores

structure for 2D analyses speeds up the calculations in comparison with full 3D calculations. The third dimension of the problem is thereby assumed to be uniform and infinite or axially symmetric. Furthermore the actual material constants for permeability etc. can be used to improve the modelling accuracy.

A. Geometry

Segmenting of the planar structure for analysis with the 2D-FEM software is performed in two different ways. By breaking the structure down into segments for which successive 2-dimensional analyses can be performed and the total losses can be obtained as the sum of the losses in the segments. The planar conductors enclosed by the core of the inductor presents little problem to 2D modelling. Modelling the conductor end-turns protruding beyond the core presents some difficulty and is approached in two ways.

In Figure 2, the conductor ends are modelled as pure half circles. With this approach the problem is defined in two parts: the main part is where the conductors are enclosed by the planar core (in x-y coordinates) and the end turns combine into a full circle (in r-z coordinates).

In Figure 3, the conductor ends are modelled as

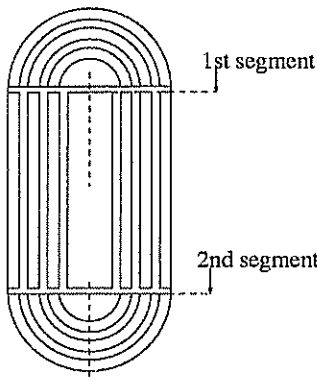


Fig. 2. Segmentation of planar structure for FEM analysis

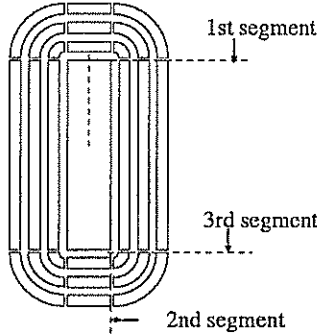


Fig. 3. Segmentation of planar structure for FEM analysis

straight sections with round corners. With this approach the problem is defined in three parts: the main part (conductors in the core) is the same as previously described while the straight parts of the conductor ends are modelled as straight conductors in air (in x-y coordinates) and the corners are defined as quarter circles (in r-z coordinates) which combine to form a circle of smaller radius than the 2-segment case.

Because of the experimental uncertainty of the air gap and the operating frequency, the FEM analysis is performed for $\pm 0.5\%$ variation of the operating frequencies and $0.4 \text{ mm} \pm 0.05 \text{ mm}$ for air gap values.

B. Materials

From in situ measurements on the windings during operation, it was determined that the conductors reached to 50°C at 20 kHz, 66°C at 40 kHz, 77°C at 60 kHz, 94°C at 80 kHz, and 102°C at 100 kHz respectively after the inverter was operating long enough to achieve thermal stability. In the theoretical calculations, the conductivity of the copper is directly related to the losses in the structure. For this reason, the values of the conductivity of copper as a function of temperature is required.

At the five temperatures, the conductivity values of copper were determined from a handbook [12]

and they were also confirmed experimentally.

Using the basic equation for resistance

$$R = \frac{l}{\sigma A} \quad (1)$$

where R is the resistance, l is the length of the conductor, σ is the conductivity of the conductor, and A is the cross sectional area of the conductor, and the equation

$$R_{t_2} = R_{t_1} [1 + \alpha_{t_1} (t_2 - t_1)] \quad (2)$$

for thermal dependency of resistance, where R_{t_2} is the resistance at temperature t_2 and R_{t_1} is the resistance at temperature t_1 and α_{t_1} is the temperature coefficient at temperature t_1 , it is possible to obtain

$$\sigma_{t_2} = \frac{\sigma_{t_1}}{1 + \alpha_{t_1} (t_2 - t_1)} \quad (3)$$

where σ_{t_1} is the conductivity value of the conductor at temperature t_1 and σ_{t_2} is the conductivity value of the conductor at temperature t_2 .

Using Equation 3 the conductivity values of copper are calculated as $5.2 \times 10^7 \text{ m}/(\Omega\text{m}^2)$, $4.9 \times 10^7 \text{ m}/(\Omega\text{m}^2)$, $4.8 \times 10^7 \text{ m}/(\Omega\text{m}^2)$, $4.5 \times 10^7 \text{ m}/(\Omega\text{m}^2)$, and $4.4 \times 10^7 \text{ m}/(\Omega\text{m}^2)$ at 50°C , 66°C , 77°C , 94°C , and 102°C .

From in situ temperature measurements on the windings and in the surrounding air during operation, it was estimated that the core reached 49°C at 20 kHz, 66°C at 40 kHz, 77°C at 60 kHz, 94°C at 80 kHz, and 102°C at 100 kHz respectively after the inverter was operating long enough to achieve thermal stability. The relative permeability (μ_r) of the core, which is 3F3 material, is determined using the databook of Philips Components. The values are taken as 2120, 2250, 2300, 2500, and 2550 at 37°C , 46°C , 52°C , 61°C , and 65°C for 20 kHz, 40 kHz, 60 kHz, 80 kHz, and 100 kHz respectively.

III. EXPERIMENTAL MEASUREMENTS

There are a number of possible methods for measuring the losses in a planar inductor experimentally. The methods are discussed with their advantages and disadvantages to justify the two methods that were used to perform the experimental measurements.

A. Loss Measurement Techniques

A.1 Voltage discharge

A series LC circuit is used in this method. A capacitor of known value is pre-charged to a determined voltage and then discharged through the series inductor while the decaying resonant transient of voltage and/or current is measured.

By analyzing the measured transient current waveform in the inductor, the power loss in the inductor can be calculated. A major shortcoming of this method is the requirement for a perfect switch.

A.2 Direct watt meter measurement

This method is very easy to use and to measure the power loss on any kind of shunt which carries the full current and voltage information. It's very difficult to find a watt meter, which can accurately measure low power loss at high frequencies.

A.3 Instantaneous power measurement

For this method a digital storage scope or waveform recorder is required. The voltage across the inductor is recorded on one channel while the current through the inductor is simultaneously recorded on another channel while the inverter is functioning. Average power loss in the inductor can then be obtained by multiplication of the instantaneous current and voltage values and taking the average over one cycle.

Although this seems easy the errors introduced by sampling and digitizing are of the same order as the power loss being measured. The method is therefore not acceptable for the current investigation. The method is suited to determining in-circuit switching losses of switching devices.

A.4 Heating of a flowing coolant

The power loss of the in situ measurement of planar inductor can be determined using a flowing liquid passing through the inductor. The temperature of the liquid on entry and exit must be determined very accurately. Using the temperature difference of the flowing liquid when it passes through the inductor, the heat dissipation in the inductor can be determined.

The flow rate of the fluid must be very slow if a low power loss measurement is targeted. The heat transferred from the planar inductor to the flowing liquid will be very slow, which makes this method impractical. This method may be practical for high power loss measurements.

A.5 Heating of a coolant in a container

In this method, the inductor is placed in a closed insulated container that is filled with a suitable coolant which is circulated to ensure uniform heat distribution. AC excitation is then applied with the converter circuit under its designed operating conditions. The temperature increase of the coolant inside the container is measured with a thermocouple and the steady state temperature is registered. Starting at the same ambient temperature, DC excitation is then applied to a resistive structure (in this case the windings of the inductor are used) placed in the same container. The DC excitation is adjusted until the same temperature of the previous AC case is registered and the DC power supplied to the resistance will then be equal to the losses in the original problem with AC excitation.

One disadvantage of this method is the long time required to reach the right temperature in the control experiment, with DC excitation. If the temperature coefficient of the coolant is high, this can become a serious problem. Another problem encountered with this method concerns the characteristics of inductor construction materials that will be placed in the coolant. Although the printed circuit board is sealed, it is still sensitive to large changes in humidity because of small damaged spots on its sides. This changes the characteristics of the structure. If a liquid coolant such as oil or water it becomes very difficult to prevent changes in the characteristics of the planar structure. Using a liquid coolant also requires a special container. In addition the mass of the liquid makes the experiment impractical. For these reasons air is used as the cooling fluid.

B. Experimental procedure

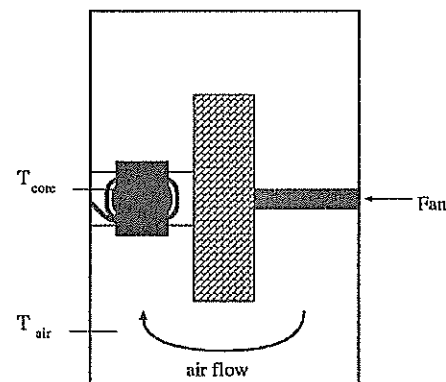


Fig. 4. Top view of the experimental setup

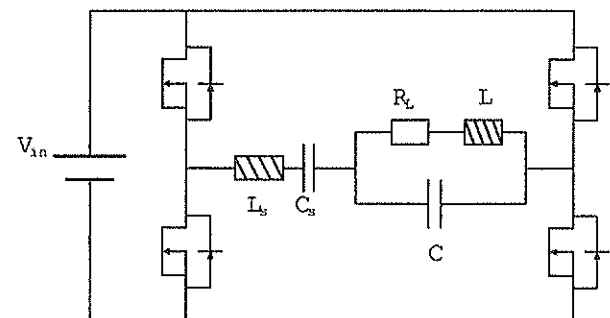


Fig. 5. DC-AC inverter circuit diagram

The advantages and disadvantages of four different methods are given. At about 5 watt loss, the accuracy of measuring the power is very important and the last method is a very good one if the air is used as a coolant in the container.

The experimental method used in this study is the same as the method explained in part III.A.5. In this case instead of a liquid, air is used as a

coolant. In order to be able to compare the temperature measurement values, the temperature of the room must be constant. This is essential because the heat flow from the inside of the container to the outside of the container must be same when the internal air temperatures of the container are the same. If the ambient temperature of the room changes, the flow rate of the heat flowing from the inside to the outside of the container will change and there will be no chance to compare the increases in the temperature between the two cases (AC and DC excitation respectively). For this reason, all measurements are performed in a constant temperature calibration room. The room temperature only varies $\pm 0.05^\circ\text{C}$ while the experiments are performed. All the measuring devices were calibrated and kept in the room throughout the experiments.

The planar inductor structure is placed in an airtight container while still maintaining its electrical contacts to the inverter circuit. With no leakage of the contained air, only heat transfer between the contained air and the container, and between the container and the air in the room is possible. To further minimise this leakage of heat the container is constructed from material with low thermal conductivity.

Using a fully functioning DC-AC resonating inverter, the inductor was excited at a selected current value. The maximum possible current is applied and this value is restricted by the saturation point of the core material and the temperature of the copper PCB tracks to prevent damage to the PCB structure through overheating. The rms value of the current applied to the inductor was controlled by means of a calibrated digital oscilloscope which had an accuracy of ± 1 per cent at the working frequencies.

In the second part of the experiment DC current is applied to the inductor, after disconnecting it from the converter. The temperature of the copper is allowed to reach the same values as with AC excitation, which are 50°C , 66°C , 77°C , 94°C , and 102°C , respectively. When the temperature has stabilized at the required value, the voltage and current values are registered to calculate the loss power and the resistance. Using equation

$$\frac{V}{I} = \frac{l}{\sigma A} \quad (4)$$

the conductivity values of the copper are determined experimentally.

The core temperatures were determined by evaluating the average temperature of the air inside the container and the copper conductor because the temperature of the core is decreasing beginning from the inner part of it to the outer part of it.

IV. EVALUATION OF CALCULATIONS AND EXPERIMENTAL RESULTS

Due to experimental difficulties with the calibration of the current measurement setup required for the described thermal experiments it was not possible to perform calculations for a single value of AC current amplitude in the inductor. For this reason the "true" operating conditions were bracketed by performing calculations for AC current amplitudes known to be larger or smaller than the actual experimental value. The FEM calculations were therefore performed for 7 A and 8 A peak values of AC current excitation.

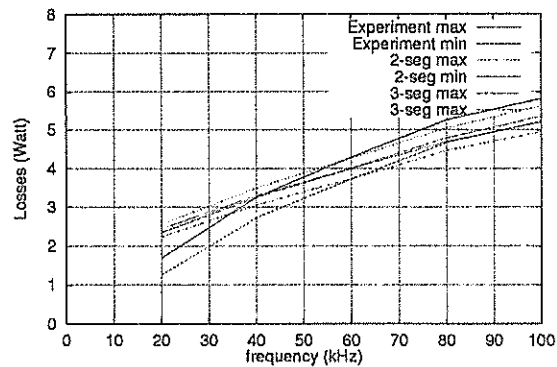


Fig. 6. Graph of experimental measurements and FEM calculations for 7 A

The experimental results and the FEM calculations for the two cases are presented in Figure 6 for the calculation based on 7 A and in Figure 7 for the calculations based on 8 A.

From inspection of the presented results it is clear that calculations tend to give too high values of losses. From a design viewpoint this is acceptable since a conservative error will be made if a design is based on these calculations.

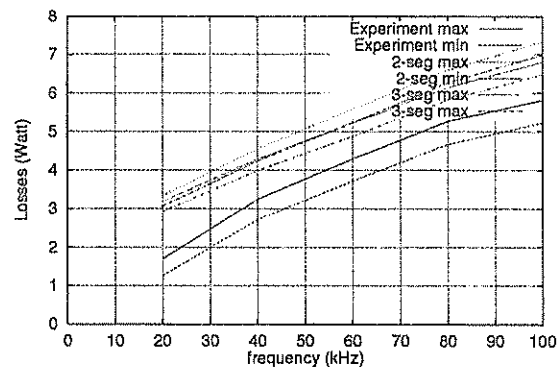


Fig. 7. Graph of experimental measurements and FEM calculations for 8 A

It is also clear from the results that the 3-segment approximation for the coil shape gives lower and

therefore more accurate values of the losses. This is what would be expected of a better approximation of the actual geometry of the conductors.

V. CONCLUSION

A large number of practical and experimental difficulties were encountered. Some uncertainties still exist regarding the accuracy of the current measurements. Nevertheless reasonable agreement between calculated and measured losses is obtained. More experimental measurements and calculations still need to be performed to ensure the overall applicability of the methods.

With the experimental infrastructure that was developed in the course of this investigation the losses in many different reactive components can now be investigated. This is especially of interest in the design of planar magnetic components for power electronic converters. At the same time the calculation methods (two-dimensional finite element method) has been proven suitable for the loss calculations during the design stage of the planar magnetic components of interest. Further applications to investigate are the use of multiple layers of conducting material sandwiched between insulating layers.

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MODELLING AND DYNAMIC COMPENSATION OF STEADY-STATE HARMONICS IN A PLASMA BURNER PLANT

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Abstract: The harmonics produced by three 120 kW plasma-burner 6-pulse converters introduced resonance into a 380 V system in a Silicon-tetra-fluoride plant that also supplied the lighting and caused circuit-breaker burnouts. With two 150 kVA dynamic compensators available, an investigation had to be carried out by modelling, to establish the suitability of these compensators for mitigating the harmonics. The paper describes suitable distortion power theory for analysing the problem. It then describes a basic steady-state frequency-domain modelling technique and proceeds to validate the model through physical measurement. Lastly, it models the system when the two compensators are installed and makes recommendations for their installation and adjustment.

Keywords: Harmonics, Compensation, Filters

1. INTRODUCTION

Although packages are available for modelling unbalanced steady-state harmonics in three-phase systems, these packages invariably have rigid structures that do not permit manipulation to suit individual needs. In collaboration with others, a new approach was made to the problem and a complete model was developed on a Mathcad Package^[1]. This model has now been further extended and refined and has been used in modelling the harmonics produced in an industrial plant for the production of Silicon-tetra-fluoride (SiF₄). At the time, this plant employed three plasma burners of 120 kW capacity each, served by 6-pulse phase-controlled AC/DC converters.

The initial problems that prompted an investigation began when it was discovered that the overload coils in the lighting distribution circuit breakers were overheating. Investigation showed that the overheating was the result of high frequency eddy-currents brought about by resonance of certain harmonics with the fluorescent lamp ballasts in the 380/220 V transformer secondary from which the burners were also drawing their power.

2. POWER DEFINITIONS

In the modelling used, it is convenient to define distorted voltage and current as complex, rms-scaled, time-dependent quantities that follow from manipulation of the complex Fourier Series^[2]:

$$u(t) = \sum_n u_n(t) \quad \text{and} \quad i(t) = \sum_n i_n(t) \quad (1) \quad (2)$$

These quantities are formed by summing the constituent n -th order time-dependent complex quantities:

$$u_n(t) = U_n e^{jn\omega t} \quad \text{and} \quad i_n(t) = I_n e^{jn\omega t} \quad (3) \quad (4)$$

over the harmonic order range.

The complex U_n and I_n respectively represent the n th order *harmonic phasor* values of voltage and current:

$$U_n = U_n e^{j\alpha_n} \quad \text{and} \quad I_n = I_n e^{j\beta_n} \quad (5) \quad (6)$$

in which α_n and β_n represent the phase angles of the voltage and current. In reality (1) and (2) represent the more general form of the scalar (and physically measurable) time-dependent voltage and current:

$$u(t) = \sqrt{2} \operatorname{Re} \left\{ \sum_n u_n(t) \right\} \quad \text{and} \quad i(t) = \sqrt{2} \operatorname{Re} \left\{ \sum_n i_n(t) \right\} \quad (7) \quad (8)$$

The *total time-dependent real power* (in all the harmonic orders) at a measuring cross-section, can be defined as:

$$p(t) = \sum_n \sqrt{2} \operatorname{Re} \{ u_n(t) \} \sqrt{2} \operatorname{Re} \{ i_n(t) \} \quad (9)$$

From which it can be readily shown that the *Total Active Power* is:

$$P = \sum_n U_n I_n \cos \phi_n \quad (10)$$

$$\text{with: } \phi_n = \alpha_n - \beta_n \quad (11)$$

The rms value of the voltage and current waveforms can be shown to be:

$$U = \sqrt{\sum_n |U_n|^2} \quad \text{and} \quad I = \sqrt{\sum_n |I_n|^2} \quad (12)(13)$$

In a novel approach by the IEEE Working Group on Non-sinusoidal Situations, under the chairmanship of A. Emanuel^[3] a set of definitions is presented for

defining distortion powers in terms of total-fundamental- and harmonic constituents

The IEEE Working Group definitions are principally adopted below, along with a number of accepted and established frequency-domain methods.

The harmonic rms-components in (12) and (13) can be separated into their fundamental and harmonic components:

$$U^2 = U_I^2 + U_H^2 \quad \text{and} \quad I^2 = I_I^2 + I_H^2 \quad (15)$$

with:

$$U_H = \sqrt{\sum_{n \neq 1} |U_n|^2} \quad \text{and} \quad I_H = \sqrt{\sum_{n \neq 1} |I_n|^2} \quad (16) \quad (17)$$

The *Total Apparent Power* S is defined as:

$$S = UI \quad (18)$$

and the *Total Nonactive Power* N as:

$$N = \sqrt{S^2 - P^2} \quad (19)$$

Alternatively, S can be defined as:

$$S^2 = (U_I I_I)^2 + (U_I I_H)^2 + (U_H I_I)^2 + (U_H I_H)^2 \quad (20)$$

and grouped as:

$$S^2 = S_I^2 + S_N^2 \quad (21)$$

In which S_I is the *Fundamental Apparent Power* and S_N the *Nonfundamental Apparent Power*. In turn:

$$S_I^2 = (U_I I_I)^2 = (U_I I_I \cos \phi_1)^2 + (U_I I_I \sin \phi_1)^2 = P_I^2 + Q_I^2 \quad (22)$$

and

$$S_N^2 = (U_I I_H)^2 + (U_H I_I)^2 + (U_H I_H)^2 \quad (23)$$

The term $(U_I I_H)$ is named the *Current Distortion Power*, $(U_H I_I)$ the *Voltage Distortion Power* and $(U_H I_H)$ the *Harmonic Apparent Power*. The latter term can, in turn, be broken down into the *Total Harmonic Active Power* P_H and the *Total Harmonic Non-active Power* N_H .

$$S_H^2 = P_H^2 + N_H^2 \quad (24)$$

in which N_H is derived analogously to N in (19).

Another useful relationship is obtained when (23) is divided by (22):

$$\left(\frac{S_N}{S_I}\right)^2 = \left(\frac{I_H}{I_I}\right)^2 + \left(\frac{U_H}{U_I}\right)^2 + \left(\frac{S_H}{S_I}\right)^2 \quad (25)$$

to yield the *Normalised Non-fundamental Distortion Power*:

Error!(26)

The three RH-terms of (26) respectively represent the *current THD*, the *Voltage THD* and the product of the previous two. The significance, applicability and utility of the different components defined above are expounded in^[3].

The single-phase relationships of (12) and (13) are extended to unbalanced three-phase networks by the following definitions that lay down the *equivalent three-phase voltages and currents*:

$$U_e = \sqrt{\frac{U_a^2 + U_b^2 + U_c^2}{3}} \quad (27)$$

and

$$I_e = \sqrt{\frac{I_a^2 + I_b^2 + I_c^2}{3}} \quad (28)$$

The identical procedures used in (12) to (26), for single-phase values of voltage, current and power, can be obtained for their *equivalent three-phase values* by replacing U and I respectively with U_e and I_e , furnishing a means by which unbalanced three phase network calculations can be carried out in systems with periodic distortion.

3. HARMONIC SUPERPOSITION MODELLING

3.1 Modelling principles

The complex compound nodal network equation for harmonic order n , of a three-phase power network with k nodes is written as:

$$[I(n)] = [Y_{bus}(n)] [U(n)] \quad (29)$$

in which $[I(n)]$ and $[U(n)]$ are column vectors of dimension k , $[Y_{bus}(n)]$ is square and represents the *nodal bus-admittance matrix* of dimension- n . The elements of $[I(n)]$ and $[U(n)]$ are sub-vectors of dimension 3 and the elements of $[Y_{bus}(n)]$ are each square sub-matrices of dimension-3. The elements of the sub-vectors are respectively formed by the per-phase harmonic phasors of voltage and current defined as:

$$U_n = U_n e^{j\alpha_n} \quad \text{and} \quad I_n = I_n e^{j\theta_n} \quad (30) \quad (31)$$

and the sub-matrices of $[Y_{bus}(n)]$ represent *compound nodal-driving-point admittances* on the main diagonal and *compound transfer admittances* on the off-diagonal locations^[4]. The three-phase unbalanced node voltages can be obtained from (29) by inversion of $[Y_{bus}(n)]$ and from there all the other quantities such as currents and powers can easily be calculated.

It must be pointed out here that the essence of the superposition modelling carried out boils down to the fact that (29) is separately solved analytically for

each discrete harmonic order- n , furnishing the harmonic phasor values of voltage in accordance with (30). The analysis begins with the fundamental harmonic order $n=1$ and proceeds through all the orders up to the highest one required, say up to the 45th order as used in the present instance. When the modelling is complete for every harmonic order, the results are summed for each node in accordance with (16) to (28).

To be able to express (29) for discrete frequencies, only steady-state operation can be modelled and all the sub-component models must be specified at each of these discrete frequencies. Generally, mutual-coupling effects can be ignored in small systems and the off-diagonal compound admittance elements in the 3×3 sub-matrices can be ignored. The three respective diagonal elements are then represented by:

$$Y_L(n) = \frac{1}{R + jn\omega L} \quad (32)$$

and

$$Y_C(n) = G + jn\omega C \quad (33)$$

respectively for the series inductive and shunt capacitive elements of each of the three phases. In these equations, the harmonic order n merely scales the inductive-reactance and capacitive susceptance linearly. This is a gross approximation but it furnishes excellent results as will be shown later.

Modelling is carried out by a manual-iterative procedure that can be summarised as follows:

- a) The transmission-component parameter values are set up for the different compound sub-models in accordance with measured values, or from the manufacturer's data.
- b) The equivalent source admittance and source fundamental current value is calculated from symmetric fault parameters supplied by the supply-authority.
- c) Converter time waveforms are measured and digitally stored for computer access. These waveforms are FFT-analysed and then characterised in time- as well as in frequency-domain form. The converter models are then adjusted in the frequency domain to yield the identical time waveforms, including unbalance where required to match observed performance.
- d) When the program is run, peak converter currents and the fundamental phase of the converter currents are iteratively adjusted until each converter achieves the specified active power. This requires pre-calculation, by the program, of both the distorted converter currents and voltage waveforms at the installed converter nodes.

- e) The converter active-power requirements and observed displacement power-factor (firing-angles) during normal operation are established
- f) Circuit changes and additions, such as the adding of passive- and dynamic compensators, at different positions in the network, again necessitate additional adjustments to converter active power levels.
- g) The model furnishes a general display section by means of which the positions for node voltages and current-flow between nodes can be arbitrarily specified. Power- and other parameter displays are then governed by these choices. The normalised harmonic magnitudes and time waveforms for the currents and voltages are furnished in graphic form for these selections.

The main model is used for the main computation and employs sub-models for sources, converters, cables, filters and compensators. Circuit-components are specified in compound, matrices. In the present report, mutual coupling is excluded between phases in the compound admittance matrices because it is so difficult to obtain and as its exclusion does not make any significant difference to the results of so limited a network. The sub-models are described in the literature^[5] and will not be further discussed here.

4. VERIFICATION OF THE MODEL

4.1 Circuit Description

Before making recommendations for mitigating the harmonics in the system, accurate modelling had to be ensured. This could only be carried out with confidence after evaluating the model against practical measurements.

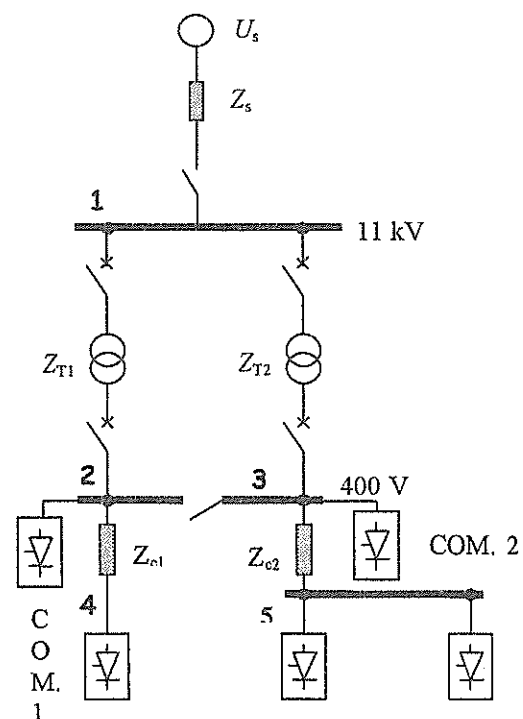


Figure 1 – Single-line circuit diagram

The diagram in Figure 1 represents the 11 kV busbars, with two 800 kVA 4.25% Impedance ΔY step-down transformers that feed the three plasma power supply converters. The bus-section between nodes 2 and 3 had to be maintained in the open position to maintain acceptable fault levels. Instead of showing the ring-feed circuit in the 11 kV supply, it is modelled here as an infinite bus with a single source impedance in an equivalent Thevenin-circuit. The rest of the diagram is self-explanatory.

The Compound Admittance Diagram in Figure 2 has been drawn directly from the one in Figure 1 and represents the circuit used in the Harmonic Superposition Model. Note that the nodes are numbered from 1 to 5. Node 1 represents the 11 kV busbars, with 2 and 3 respectively the two 380 V ones. The two transformers are modelled respectively between nodes 1 and 2 and nodes 1 and 3 as π -models. The two cables from the substation, respectively to burner-1 and to burners 2 and 3 are modelled as equivalent L-admittances. The two converters 2 and 3 are shown as connected to the same node- number 5. In this diagram, the Supply-Authority Equivalent source is modelled as a Norton Equivalent circuit.

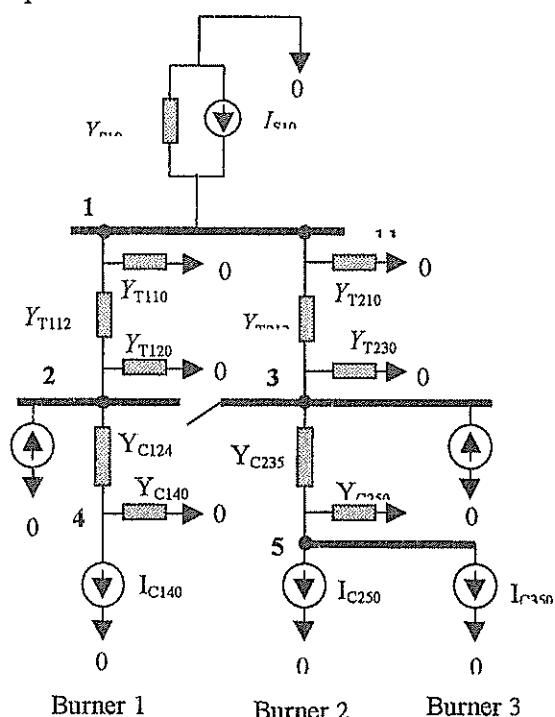


Figure 2 – Compound admittance diagram

An extensive number of measurements were carried out in which phase currents and line-line and line-neutral voltages were recorded digitally. Unfortunately the limited scope of this paper will only permit a limited number of these waveforms to be presented here, but the author will gladly discuss details of the main report on which this paper is based with interested readers. What these measurements showed, however, was that voltage distortion came into the transformer secondaries almost exclusively as a result of the relatively high transformer leakage inductances and that the source and other transmission components contributed very little.

4.2 Measurement for model verification

All three burners were operated at full power. The line-neutral voltage on node-2 and the low-voltage line-current on transformer 1 was recorded and is shown in Figure 3. The load current of burner 1 shows the characteristic 6-pulse phase-controlled converter waveform with large DC inductance and continuous output current. It also shows asymmetric triggering between the three phases in the asymmetry in the forward- as well as in the reverse current flow. Accurate simulation of such a phenomenon calls for intricate model adjustments. The notching in the voltage waveform is also clearly discernible and was the chief source of excitation for resonating with the fluorescent lamp ballasts mentioned earlier.

One of the features built into the model is that the time-domain oscilloscope measurement data can be read into it. This data is then converted to frequency-domain data and can be manipulated and analysed, just as the modelling data is.

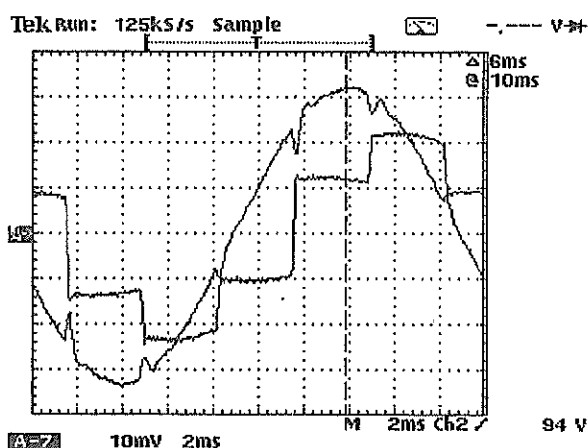


Figure 3 – Measured voltage and current node-2

Among other, the origin of all the waveforms in the measurement, as well as the vertical scale of the individual traces, can be shifted at will without

altering the waveforms to correspond, for example, with the modelled data. The measurement data from Figure 3 was manipulated in this way and is presented in Figure 4.

The model was set up with the configuration shown in Figure 2 with parameters corresponding to those in the physical network. Modelling results, corresponding to those shown in Figure 3 and Figure 4 were re-constructed in the time-domain and is shown in Figure 5. The origin on the time-axis has been shifted for the waveforms in Figure 4 to correspond with the measured results in Figure 5 to prepare the two sets of waveforms for easy comparison by inspection. The comparison shows that virtually all the features of both the voltage and the current correspond. The only superficial aberration between the two sets of waveforms is in the "shoulders" of the current, showing that the match in the unbalance between the three phases is not 100% accurate.

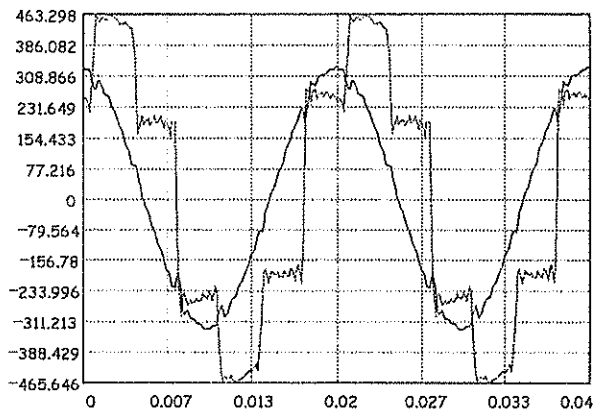


Figure 4 – Time-domain reconstruction of measured voltage and current on node-2

Even this aspect can be matched by trimming the model, but the error is negligible with respect to the task at hand, and exact trimming will require additional time and effort that is not warranted here.

4.3 Comparison of modelled results

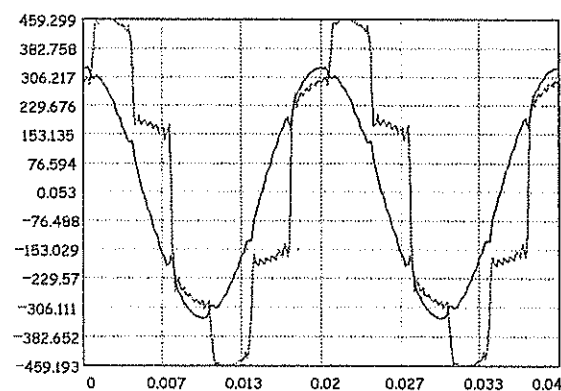


Figure 5 - Time-domain modelling of measured voltage and current on node-2

The accuracy of modelling can be further illustrated by listing a number of the numerical results for modelling against that of measurement, as in Table 1.

Table 1 – Modelled against measured results

Voltage Node:	2	2
Start Current Node:	2	2
End Current Node:	4	4
Node Voltage	231.734	227.4
Fundamental Voltage(V):	231.486	227.1325
Harmonic Voltage(V):	10.7174	11.0263
Node-node Current(A):	311.9664	313.285
Fundamental Current(A):	303.1014	300.7829
Harmonic Current(A):	73.8412	87.6192
Apparent Power(kVA):	216.8796	213.723
Non-active Power(kVA):	153.6933	150.9863
Fun. App. Power(kVA):	208.55	204.95
Current Dist. Power(kVA):	51.2796	59.7035
Voltage Dist. Power(kVA):	9.7454	9.9495
Joint Harm. Power(VA):	2.254.26	1926.07
Harm. Ap. Power(kVA):	2.3742	2.8983
Current THD(%):	24.3619	29.1304
Voltage THD(%):	4.6298	4.8546
Nor. Har. Ap. Power(%):	1.1279	1.4142
Disp. Power Factor:	0.7338	0.7413
Overall Power Factor:	0.7056	0.7078

Even here, the correspondence is remarkable and shows the accuracy of the model. The information in Table 1 also illustrates the utility of the IEEE Working-Group definitions.

5. MODELLING WITH DYNAMIC COMPENSATORS INSTALLED

The business-unit responsible for the Silicon-tetra-fluoride plant had access to two 150 kVA dynamic compensators (or Active Filters as they are sometimes erroneously referred to) from Equipaggiamenti Elettronica Industriali (EEI) in Vicenza. These two units were originally purchased for the Laser Isotope Enrichment programme that was closed down. They are designed for 3-phase operation and employ Pulse-Width Modulation at about 5 kHz through a simple control system that causes them to inject the error current between perfect sine-waves and the measured load current into the system. This control system also permits the fundamental and harmonic components to be collectively or separately compensated so that the one can be traded off against the other when the demand exceeds their capacity.

It was proposed to connect the compensators respectively to nodes 2 and 3 as shown in Figure 1 and their models are shown at these positions in Figure 2 in the form of pure current-sources. Unfortunately no frequency-response characteristics were available for the two compensators except a set of data which was compiled earlier when one of these units were installed in the original Isotope separation plant and when the model could be aligned with measured results.

To construct a frequency-domain model for the EEI dynamic compensators, certain assumptions had to be made. They are:

1. That all waveforms were periodic and representable in the frequency domain by discrete spectra.
2. That the dynamic compensators only function at these discrete spectra.
3. That the response at each discrete harmonic order can be weighed to match the dynamic response of the compensator at the relevant frequencies.
4. That the "effectiveness" of the compensators for compensating at these higher frequencies will reduce because of the reduced "bandwidth" of PWM operation as frequency increases.

To model the dynamic compensator, a table had to be compiled that set out the "weights" representative of its response for each discrete harmonic frequency as shown in Table 2. The top and third rows show the harmonic order number and the second and fourth rows the weighting factor for adjustment of the harmonic compensating current at that harmonic order. The modelled PWM-current is output as the product of the measured error-current and the weighting factor.

Table 2 -Compensator model response weighting

1	2	3	4	5	6	7	8	9	10
100	100	100	98	97	96	95	90	85	80
11	12	13	14	15	16	17	18	19	20
76	74	72	70	68	64	62	60	58	56

Analysis by means of the model showed that compensator 1, required to compensate a single burner could compensate for both the fundamental power factor and for the harmonics. Compensator 2, however had to be adjusted to compensate for the harmonics only on account of the higher load. Because of the relatively high leakage reactance of the transformers each LV section had to contend with its own harmonics and the voltage distortion on

the HV side even with all the burners operating was almost negligible.

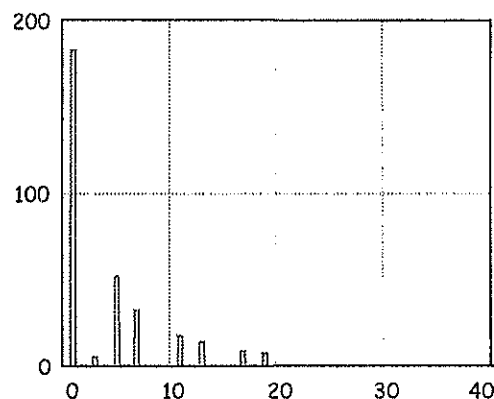


Figure 6 - Modelled current spectra for compensator 1

With the two compensators set in this manner, their modelled current spectra are as shown in Figure 6 and in Figure 7 respectively.

As shown, compensator 1 will sink the fundamental component necessary for displacement power factor correction of nearly 200 A in addition to the 5th, 7th, 11th, 13th, 17th and 19th harmonics. Compensator 2 will only handle the harmonics. Notice the presence of a small 3rd harmonic component in each. This 3rd harmonic is there as a result of the current unbalance built into the model to match the asymmetry in the firing angles of the phase controllers that have been observed previously.

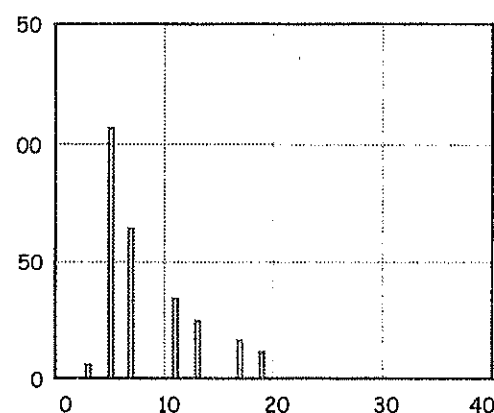


Figure 7 - Modelled current spectra for compensator 2

It is more academically instructive to display the line-currents that the two compensators will respectively draw than its utility value, but it furnishes insight and enhances the credentials of the model. The line currents for compensator 1 and 2 are respectively shown in Figure 8 and in Figure 9.

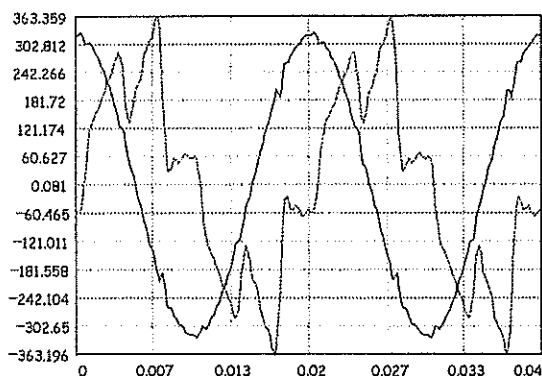


Figure 8 – Compensator 1 current and node 2 voltage

Notice the fundamental component present in the number 1 compensator current. This component is absent in the current of compensator 2 which is adjusted to compensator for harmonics alone. There are also perceptible improvements in the voltage traces on node 2 when compared with the previous results without compensation as depicted in Figure 5.

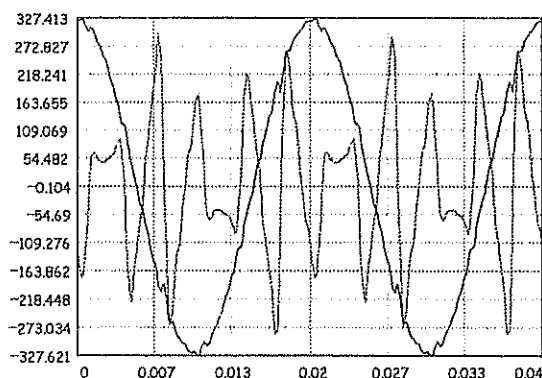


Figure 9 – Compensator 2 current and node 3 voltage

The compensator performance is best illustrated in tabular form, analysed in terms of the IEEE Working-group definitions.

Table 3 – Operating data for the compensators

PARAMETER	COM 1	COM 2
Node Voltage(V)	226.944	226.94
Fund. Voltage(V)	226.8533	226.85
Harm. Voltage(V)	6.427	6.427
Current(A)	219.78	131.212
Fund. Current(A)	209.9306	0
Harm. Current(A)	65.0688	131.212
App. Power(kVA)	141.6359	89.3337
Non-act. Power(kVA)	149.6056	89.3337
Fund. Ap. Power(kVA)	141.49	0

Cur. Dis. Power(kVA)	44.2832	89.2979
Vol. Dis. Power(kVA)	4.0477	0
Har. App. Power(kVA)	1.2546	2.5299
Voltage THD(%)	2.8331	2.8331
Disp. Pow. Factor:	0.0213	0

Notice that the limiting factor on the application of compensator 1 is in its power of 141.63 kVA compared to its rating of 150 kVA as well as in its current rating of 219.78 A, compared to its rated current of 227.9 A. In the case of compensator 2, it only draws 89.33 kVA, but its current is at 131 A. If it was possible to adjust this compensator to put all of its reserve into the fundamental correction, this would be recommended and this correction would be substantial though not to unity.

The observant reader will remark that the sum of the fundamental current of 209.9306 and the harmonic current of 65.0688 does not add up to the total current of 219.78 for compensator 1 and he will be right. If he adds the squares of these two currents and draws the square-root he will, however, come to the exact figure. The reason for that is that the harmonic current is orthogonal^[9] to the fundamental current.

6. CONCLUSIONS AND RECOMMENDATIONS

The author believes that the above exercises have shown the utility of harmonic superposition (frequency-domain) modelling of networks with distortion loads. The contents of this paper just present a number of basic excerpts from the main report and is far from complete in that respect.

Unfortunately the compensators could not be installed and their performance monitored to further validate the accuracy of the modelling. That will have to wait for later and for a later publication.

The utility and simplicity of the IEEE Working-Group definitions also emerged from the analysis and it is recommended that these definitions be adopted in this type of work generally.

7. ACKNOWLEDGEMENT

The author acknowledges the leniency granted him by the Atomic Energy Corporation of SA for publishing this work and in that regard specifically thank Messrs Retief and Martens for their assistance in the measurements and in acquiring the relevant distribution system data needed.

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EDDY CURRENT SHIELDING AND INDUCTIVE FORCE ANALYSIS OF PARAMAGNETIC SHIELDS IN LINEAR MOTORS

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Abstract: A small scale prototype of a linear synchronous motor has been designed, built and tested. Present studies include the investigation of the shielding properties of a paramagnetic material as well as its use as a stabiliser during asynchronous operation. Analysis has shown that a shield made of aluminium reduces eddy currents in the permanent magnets and back iron of the secondary of the linear synchronous motor. An optimised plate thickness of 3 mm has also been ascertained. This optimised plate thickness gives a maximum static inductive thrust force of 744 N (unscaled). A copper shield of similar dimensions gives a maximum static inductive thrust force of 945 N (unscaled). Aluminium is best suited as a shield owing to its superior conductivity, lower density and lower cost than that of copper. The shield performs more of a damping action rather than a lifting action.

Keywords: Eddy currents, shielding, inductive forces.

1. INTRODUCTION

This study involves the investigation of the use of a paramagnetic material, such as aluminium, to shield the permanent magnets and back iron from eddy currents in the secondaries of linear synchronous motors (LSMs).

The LSM secondary that will be discussed is that of the surface mounted design owing to its superior static thrust force and dynamic braking characteristics [1].

This study is divided into two broad categories, each having its own subdivisions. The first category deals with the magnetic shielding properties of the paramagnetic material. The second section deals with the inductive forces that are induced in the shield during asynchronous operation.

A cross-sectional view through the secondary of the surface mounted secondary is shown Figure 1.

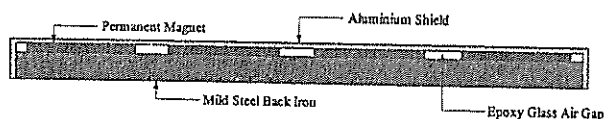


Figure 1 Cross section of LSM surface mounted secondary

2. EDDY CURRENT INVESTIGATION

There are numerous tools available in order to investigate the phenomena of eddy currents. Various methods were used in conjunction with each other in order to give a fuller understanding of the problem at hand.

The various methods that will be discussed will include empirical analysis, numerical analysis concluding with the finite element method (FEM) of analysis. These methods are then verified by physical testing.

2.1 Empirical Analysis

The empirical analysis and initial starting point involved the following standard equation for eddy current joule loss :

$$P_e = \frac{V \cdot \pi^2 \cdot f^2 \cdot \tau^2 \cdot B_m^2}{6 \cdot \rho} \quad (1)$$

where:

P_e the eddy current loss (J)

V is the volume of material (m^3)

f is the frequency (Hz)

τ is the material thickness (m)

B_m is the maximum magnetic flux density (T)

ρ is the resistivity of the material (Ωm)

The use of this equation leads to the following graphical representation. Two cases are plotted in Figure 2. One case is when there is no shield present and the other is with a 2 mm aluminium shield. Reduction in losses can be clearly seen when the 2 mm aluminium shield is present.

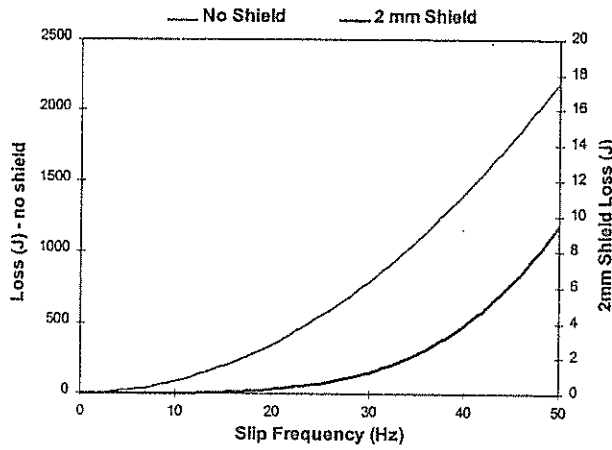


Figure 2 Empirical eddy current loss curves

2.2 Numerical Analysis

The numerical method used is that developed by Siakavellas [2]. An average air-gap flux density of 0.2 T is assumed (verified using FEM). A sinusoidal flux density waveform is also assumed.

The analysis calculates the current densities, J_x and J_y , in the plates in the x- and y- directions as a function of time. The circulating current and power dissipated can also be determined.

The equations used are as follows :

$$J_x(y) = \sigma \dot{B} \frac{l^2}{l^2 + a^2} y \quad (2)$$

$$J_y(x) = -\sigma \dot{B} \frac{a^2}{l^2 + a^2} x \quad (3)$$

$$I(t) = -\frac{1}{8} \sigma h \frac{l^2 a^2}{l^2 + a^2} \dot{B}(t) \quad (4)$$

$$P(t) = \frac{1}{16} \sigma h \frac{l^3 a^3}{l^2 + a^2} \dot{B}^2(t) \quad (5)$$

where :

σ is the conductivity ($\Omega^{-1} \text{m}^{-1}$)

$\dot{B}$ is the derivative of the magnetic field $B(t)$

l is the length of the plate (m)

a is the width of the plate (m)

$I(t)$ is the circulating current waveform (A)

$P(t)$ is the dissipated power waveform (W)

The following pair of figures shows the current densities in the x- and y- directions as a function of time. The plate material is aluminium and is 2 mm thick.

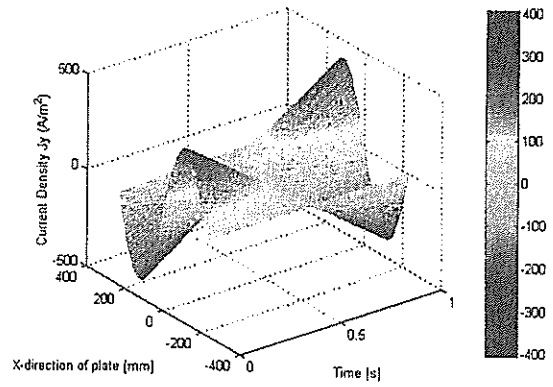


Figure 3 J_y current density at 1.0 Hz slip frequency

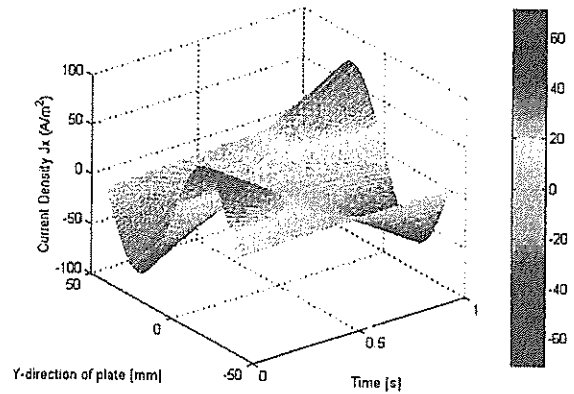


Figure 4 J_x current density at 1.0 Hz slip frequency

2.3 FEM Analysis

Two models of the LSM were investigated using FEM analysis. The only difference between the two models, is one has its secondary shielded by aluminium and the other is unshielded.

The following figures, (Figure 5 and Figure 6), are plots of the eddy current contours.

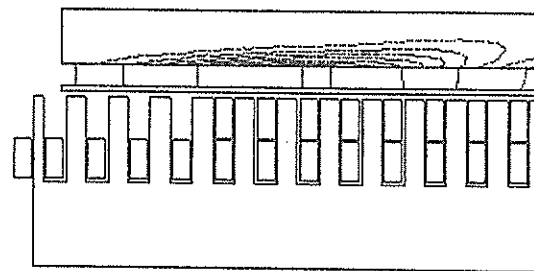


Figure 5 System without aluminium shield showing eddy current flow in the back iron

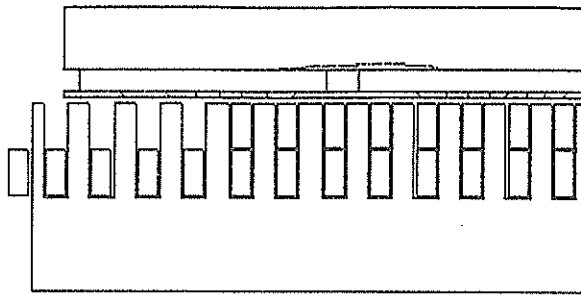


Figure 6 System with aluminium shield showing eddy current flow only in shield

Figure 5 clearly shows the contoured nature of eddy currents in the back iron, whereas in Figure 6, there is only eddy current flow in the aluminium.

These two figures are a cross-sectional view of the eddy currents that are flowing in the z-plane (that is, into the paper). These simulations were performed at a slip frequency of 5 Hz, thus skin effects can be ignored

2.4 Physical Testing

Physical testing is performed by placing search coils over the magnet faces and running the LSM with and without the aluminium cover. Heat runs were also performed to evaluate the difference in joule loss between having the shield and not having the shield.

3. INDUCTIVE FORCES

It is well known from the theory and testing of linear induction motors (LIM) that there exists an inductive thrust force. These forces were investigated, first using a LIM and then on the LSM prototype.

3.1 LIM Analysis

The LIM was investigated in order to ascertain scaling factors between FEM results and those achieved in reality. The FEM package being used was only two-dimensional and assumed a unit depth into the plane of 1m. Owing to the path of the eddy currents the results obtained in FEM could not simply be linearly scaled to the physical system that was 76.2 mm deep.

The LIM was also investigated to ascertain the shield thickness that would give a maximum inductive force. The maximum forces for different plate thickness are presented in Figure 7.

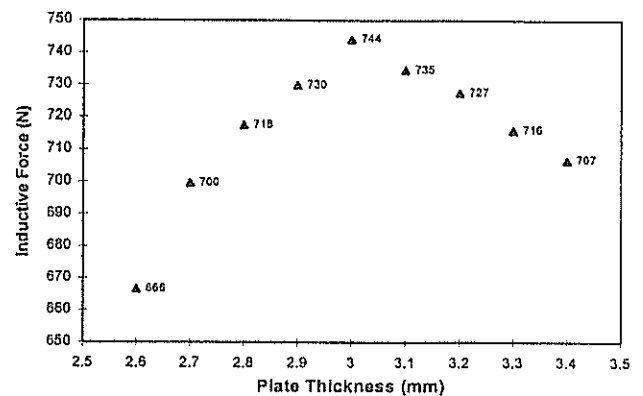


Figure 7 Comparison of inductive forces with varying aluminium plate thickness'

For the comparison to be correct for the different plate thickness', the flux density in the air gap was kept constant at 0.2 T. It can be seen from this figure that an aluminium plate 3 mm thick, gives the maximum inductive thrust force. It must be kept in mind that these values are not scaled.

A physical model of the LIM used in the FEM domain was tested to ascertain a scaling factor. The LIM was excited with the same parameters as that used in the FEM domain. The following graph depicts the results.

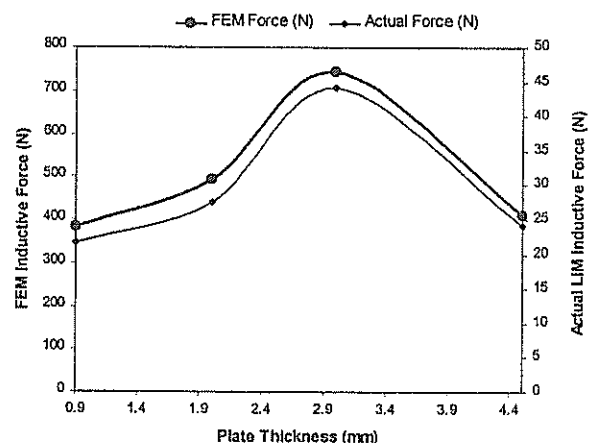


Figure 8 Comparison between FEM and physical testing on a LIM

From the testing that was performed a scaling factor of approximately 17.5 is obtained. The concept of the scaling factor means that if the FEM result is divided by 17.5 then the actual physical result would be obtained. This scaling factor is only correct for these particular systems and is not a global scaling factor.

For completeness, the same analysis was performed on a LIM having a copper cap instead of an aluminium one. The results lead to the discovery that the copper

cap provides a greater inductive force than compared to a similar aluminium cap. The case for 2 mm aluminium and 2 mm copper caps is plotted in Figure 9.

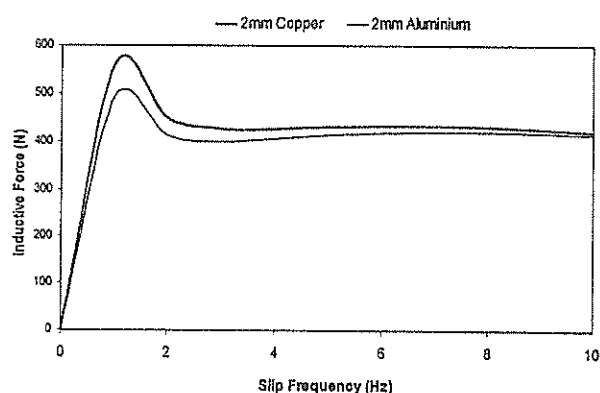


Figure 9 Comparison of inductive forces between a 2 mm aluminium and 2 mm copper cap

3.2 LSM Inductive Forces

An aluminium shield is investigated as it has a lower density than copper as well as lower cost.

The analysis that was performed for the LIM was also carried out on the LSM. FEM simulations were run for a 2 mm thick shield. Simulations were also performed for a shield 3 mm thick as this is the shield that gives the greatest static inductive force.

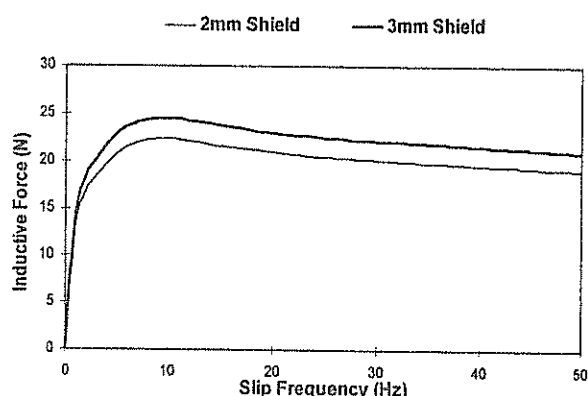


Figure 10 LSM inductive force curves for 2 mm and 3 mm aluminium shields

When the scaling factor is used, an inductive force of 42.5 N is obtained. As there are two aluminium shields on the LSM, this corresponds to a maximum scaled inductive force of 85 N. The maximum static thrust force of the surface mounted LSM was found to be 1535 N [1]. Thus the inductive force corresponds to only 5.5 % to that of the main synchronous force

4. CONCLUSIONS

It can be concluded that the incorporation of a paramagnetic shield around the secondaries of linear synchronous motors reduces eddy current loss within the permanent magnets and mild steel back iron as well as producing inductive forces during start-up and transients.

The optimum plate material would be copper but its cost and density make it prohibitive. A 3 mm aluminium shield was found to be the most suitable. The inductive force generated by this 3 mm aluminium shield was 744 N (unscaled) or 43 N (scaled), per side. This force was only 5.5 % of the main synchronous force, hence the shields perform more of a damping function than a lifting function.

These results have been obtained through various methods such as empirical analysis, numerical analysis, finite element method of analysis and physical testing. There is good correlation between the results obtained theoretically as compared to those obtained practically.

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MOTOR LOADING CAUSING FLICKER AT VARIOUS FLICKER DISTURBING FREQUENCIES

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Abstract:

This paper investigates the design and operation of a flicker generating motor load, operating a various flicker-disturbing frequencies and at different values of resistive load.

Keywords:

Voltage flicker, Flicker emission, Motor loads

1. INTRODUCTION

A lot of AC powered electrical equipment incorporate automatic switching devices. The current drawn by this equipment varies as different loads are connected across the supply. This variation in current will cause variations in supply voltage. Other devices connected to the same supply will experience this supply voltage fluctuation. This affects the operation of these devices [1]. The severity of the voltage fluctuations is related to the size of the changing load in relation to the strength of the power supply. A fluctuation can be defined as the difference between the loaded and unloaded voltages, which is also the voltage drop in the power delivery system [2].

Electrical utilities normally operate their generators at constant mechanical speed and constant electrical voltage. The quality of service of the supply utility is

determined by the degree to which these two quantities are maintained constant.

2. CHARACTERISTICS OF POWER SYSTEMS

Changes in the mechanical torque and in the electrical currents on the system cause the supply generators to deviate from normal mechanical speed and electric voltage. After a disturbance, the balance of the system can be restored to obtain normal speed and voltage by adjusting the input to the generator's prime mover, adjusting the excitation of synchronous machines, or adjusting the transformer ratios used in the system [3].

Cases of voltage flicker rarely originate in the power supply system and are eliminated as soon as possible when it is detected. Most of the flicker on central distribution systems is due to the customer's utilization equipment [3].

3. CAUSES OF VOLTAGE DIPS AND FLICKER IN POWER SYSTEMS

A certain amount of voltage fluctuation must be tolerated in the supply system because the system cannot compensate for rapid load variations. The voltage fluctuations cannot be accepted when malfunctions, deterioration of connected equipment or intolerable flicker occur [4].

Utilities make large capital investments to obtain equipment that are designed for automatic control of speed and voltage. All such equipment has an inherent time-delay characteristic. The regulating devices must detect the need to correct the power system when a deviation is made from normal operational voltage and speed. The corrective action must be damped, or delayed enough, to prevent oscillations and the magnitude of the controlled quantity is held within a range of minimum and maximum values when relatively slow changes of loading occurs [3].

When the electrical current or mechanical torque changes suddenly, the deviation from the normal operating voltage, or speed, must be outside the bandwidth of normal regulation before a correction can be made to restore these operating values to acceptable limits. A voltage dip is an occasional single disturbance caused by short circuits, faults, and inrush currents during motor starting. Voltage flicker can be described as a series of voltage dips [3].

4. VOLTAGE FLICKER

The IEC standards define flicker as the influence that an unsteady light intensity has on the visual sensation experienced by humans. Flicker is dependent on the spectral distribution of the fluctuating luminance. Flicker quantifies a subjective process, which equates measured variations in the main supply voltage to perceived changes in the brightness of a lamp [1].

Voltage flicker refers to a slow (0,5 Hz to 30 Hz) modulation of the voltage waveform. This periodic fluctuation can be described as an amplitude modulated signal, with the fundamental power frequency as the carrier signal and the flicker voltage as the modulating signal. This flicker voltage magnitude is called the instantaneous flicker level. The characteristics of the instantaneous flicker level depend on the size and type of the load that is producing the disturbance [5].

Any large pulsating load that is connected to the power system will cause the supply voltage to fluctuate, and will produce flicker if the frequency of the fluctuation is within the flicker frequency range of 0,5 Hz to 30 Hz [2].

Small frequent changes in the supply voltage, or large infrequent changes in the supply voltage, may not cause flicker problems; while large voltage variations that occurs frequently will cause flicker [1]. Cyclic or rapidly recurring voltage changes are generally more objectionable than non-cyclic voltage changes [3].

Motor-driven reciprocating loads, such as compressors, produce cyclic pulsating flicker voltages. The comparatively small variations in the supply voltage may be objectionable if the voltage pulsates between 6 to 12 times per second [3].

Fluctuations of this type are revealed by variations in the luminous flux of incandescent and fluorescent lamps, causing a disagreeable physiological sensation to those who are subjected to the disturbed illumination. These fluctuations can disturb the proper operation of electronic equipment of all types [6].

The strongest human response to flicker occurs when luminance changes occur between 800 and 1200 changes per minute, which is 400 to 600 cycles per

minute or 7 Hz - 10 Hz. The human visual system will be able to perceive voltage fluctuations of 0,3 %, and above, as flicker if the voltage fluctuations occur in this frequency range. The human threshold to flicker perception, and the irritation of the flicker perceived, are very dependent on the observer of the flicker [2].

5. TOLERANCE LIMITS FOR VOLTAGE DIPS AND FLICKER

The permissible amount of flicker voltage on a supply system cannot be stated concisely for a number of reasons:

- Flicker is primarily objectionable because of the reaction of the human eye to variations in electric light intensity. The reaction of different individuals varies over a wide range and flicker that is objectionable to one person may not even be perceptible to another person.
- Different sizes and types of electrical lamps produce different levels of luminance changes when subjected to the same voltage flicker.
- The magnitude of the voltage change, the rate at which the voltage changes, the duration of the voltage change, and the occurrence frequency of the voltage changes effect the limit of flicker tolerance [3].

6. DESIGN OF A FLICKER LOAD

Power semiconductor devices such as diodes, thyristors, triacs and power transistors are used in power applications as switching devices. Many applications of the semiconductor devices are found where a DC supply is used. The load power of a DC system can be controlled by switching the load on and off. Circuits using this technology are referred to as DC choppers [7].

The purpose of this study was to design a thyristor switching circuit that will switch a resistive load at different frequencies onto a DC generator.

A chopper circuit employing parallel-capacitor commutation was used for switching the resistive load onto the generator. An op-amp pulse triggering circuit provided the thyristor gate triggering pulses. Power transistors amplified the output of the op-amp pulse triggering circuit, before it was fed to the thyristor gates.

The op-amp circuit provided gate-triggering pulses in the following pre-selectable frequencies: 1 Hz, 3 Hz, 5 Hz, 6 Hz, 7 Hz, 8 Hz, 10 Hz, 15 Hz, 20 Hz and 25 Hz. The frequencies were calibrated with an oscilloscope and were concentrated around the 7 Hz region because the strongest human response to flicker occurs between 7 Hz to 10 Hz [2].

Block Pulse firing was used to trigger the thyristors, shown in the circuit diagram of figure 1.

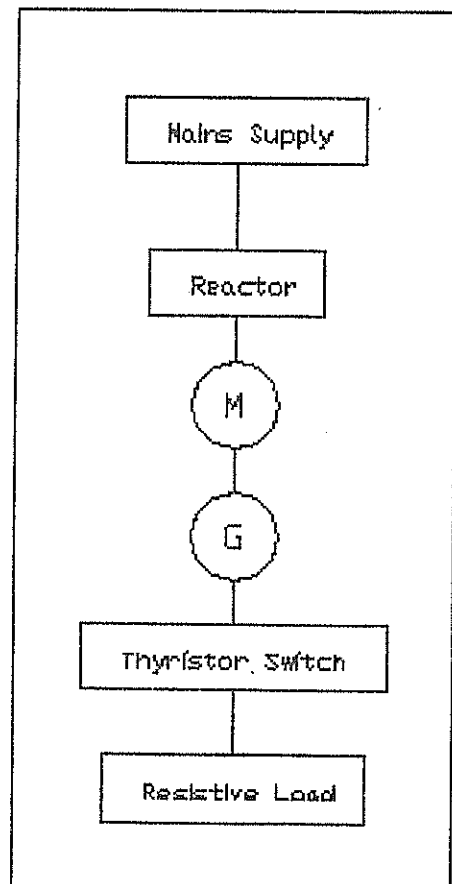


Figure 1: Single line diagram of test system.

7. RESULTS OF FLICKER LOAD TEST CARRIED OUT ON AN INDUCTION MOTOR

A dc generator, which was driven mechanically by a 3-phase induction motor, was loaded with resistive load switched at predetermined frequencies in order to produce cyclic pulsating load on the induction motor. The tests were carried out for different load currents.

7.1 Test results at Load current = 20 A

The first test was conducted by regulating the reactor output voltage at 380 V and the DC generator's voltage at 225 V. The current of the generator was regulated to 20 A by adjusting the resistive load. The resistive load was 4,5 kW. The flicker measurement results obtained during the test is shown in table 1:

Table 1: Flicker and flicker emission contribution at load current of 20A

Pulsating load frequency	Flicker at Input of Reactor	Flicker at Output of Reactor	Flicker Emission Contribution
1Hz	2,02	2,13	0,63
3Hz	2,16	2,39	0,82
6Hz	2,39	3,15	1,74
8Hz	2,51	3,77	2,18
10Hz	1,57	2,41	1,44
15Hz	0,69	1,07	0,66
25Hz	0,46	0,52	0,20

Flicker emission was measured across the reactor and used in equation (1) to determine the actual flicker contribution of the flicker load:

$$Pst = Pst_{Emission} \frac{S_{CCR}}{S_{CCN}} \quad (1)$$

Where: S_{ccn} = Short circuit power of the network
 S_{ccr} = Short circuit power of the reactor

S_{ccn} was 2,78 MVA and S_{ccr} was 4,30 MVA during the tests. The graphical representation of the flicker and flicker emission contribution, which was measured when the load current was 20 A, is shown in figure 2:

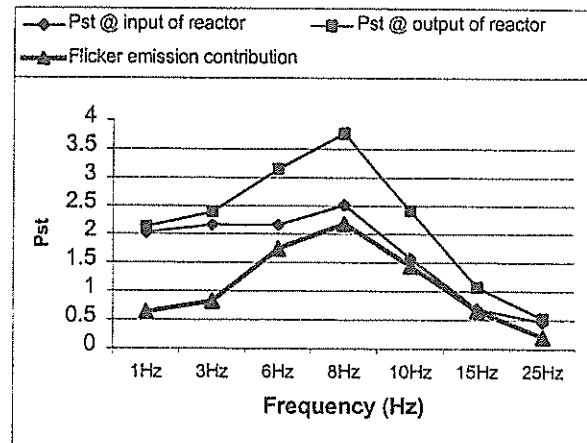


Figure 2: Graphical representation of flicker and flicker emission contribution at load current of 20,5A

7.2 Test results at Load current = 31,5A

The second test was conducted by regulating the reactor output voltage at 380 V and the DC generator's voltage at 230,5 V. The current of the generator was regulated to 30,5 A by adjusting the resistive load. The resistive load was 7,26 kW. The flicker measurement results obtained during the test is shown in table 2:

Table 2: Flicker and flicker emission contribution at load current of 31,5A

Pulsating load frequency	Flicker at Input of Reactor	Flicker at Output of Reactor	Flicker Emission Contribution
1Hz	1,04	1,69	1,03
3Hz	1,33	2,2	1,36
6Hz	2,63	4,26	2,59
8Hz	3,21	5,29	3,21
10Hz	1,98	3,29	2,05
15Hz	0,93	1,48	0,90
25Hz	0,32	0,49	0,29

The graphical representation of the flicker and flicker emission contribution, which was measured when the load current was 30,5 A, is shown in figure 3:

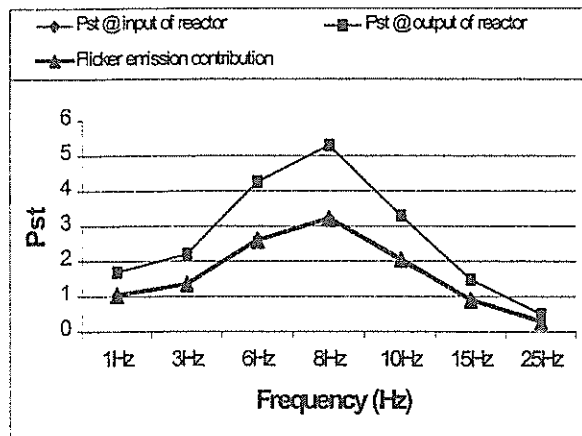


Figure 3: Graphical representation of flicker and flicker emission contribution at load current of 31,5A

7.3 Test results at Load current = 40,5A

The third test was conducted by regulating the reactor output voltage at 380 V and the DC generator's voltage at 230 V. The current of the generator was regulated to 40,5 A by adjusting the resistive load. The resistive load was 9,32 kW. The flicker measurement results obtained during the test is shown in table 3:

Table 3: Flicker and flicker emission contribution at load current of 40,5A

Pulsating load frequency	Flicker at Input of Reactor	Flicker at Output of Reactor	Flicker Emission Contribution
1Hz	1,33	2,15	1,27
3Hz	1,78	2,89	1,68
6Hz	3,26	5,6	3,46
8Hz	3,98	6,73	4,29
10Hz	2,63	4,45	2,85
15Hz	1,26	2,01	1,25
25Hz	0,63	0,8	0,40

The graphical representation of the flicker and flicker emission contribution, which was measured when the load current was 41,5 A, is shown in figure 4:

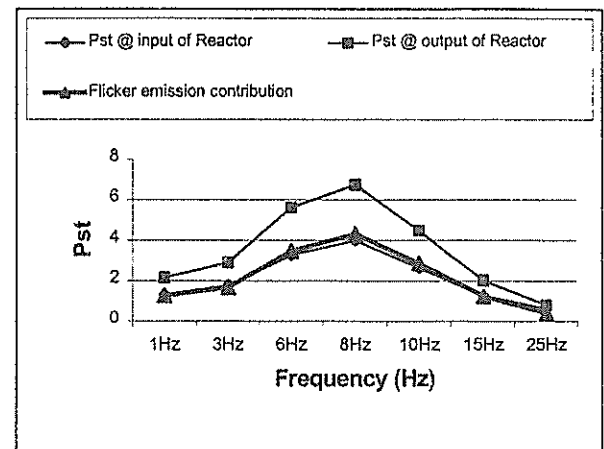


Figure 4: Graphical representation of flicker and flicker emission contribution at load current of 40,5A

7.4 Comparison of flicker emission tests

The flicker emission contribution results obtained in the study is correlated in figure 5:

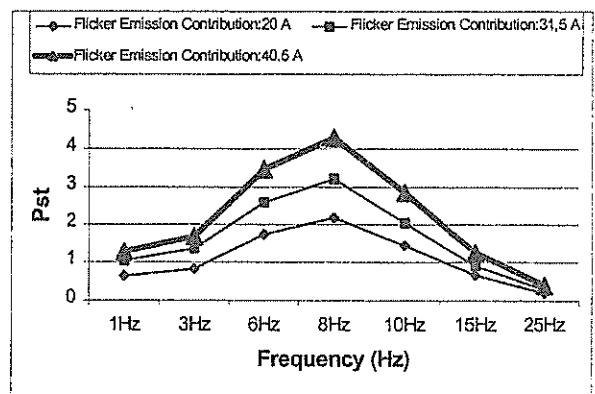


Figure 5: Graphical correlation analysis of flicker emission contribution levels measured at different loads

The flicker emission contribution of the flicker generating motor load clearly increased as the resistive load connected to the DC generator increased.

8. CONCLUSION

The utility must provide their flicker-producing customer with a supply that has a sufficiently large short circuit capacity to limit flicker emission. The

magnitude of the voltage disturbance is dependent upon the ratio of the current causing the disturbance to the short-circuit capacity on the bus where the disturbance is observed. The fundamental solution must be to hold the supply voltage within tolerance limits.

The strength of the supply should not be determined by the average operating power demand, but by the necessity to withstand reactive power swings.

The relative voltage change ΔV , which is associated with a load change, is evaluated by the relationship between the load change and the short circuit power S_{SC} of the supply network. The relative voltage changes of the phase to neutral voltages, and of the phase to phase voltages, are given in equation (2) for balanced three phase loads:

$$\Delta V = \frac{\Delta Q}{S_{SC}} + \left[\frac{\Delta P}{S_{SC}} * \frac{R}{X} \right] \quad (2)$$

Where: ΔP = Change in active power of the load
[kW]

ΔQ = Change in reactive power of the load
[kVAr]

S_{SC} = Short circuit power (fault level) of the system [MVA]

R/X = Ratio of the supply resistance to reactance [8]

The flicker motor load operated successfully and illustrated how the flicker level is influenced by the magnitude of the load.

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LINEAR SYNCHRONOUS MOTOR HOISTING SYSTEMS FOR ULTRA-DEEP LEVEL MINING

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Abstract - This paper deals with linear synchronous motors and their suitability as viable ropeless alternatives to conventional hoisting technologies in ultra deep level gold mining. This technology is extended to an integrated transport system throughout the mine. A brief overview of the ongoing research being conducted at the Department of Electrical Engineering, University of the Witwatersrand is given. Transport requirements around the mine are defined and limitations of conventional systems are listed. A long primary, short secondary linear synchronous motor is introduced with possible advantages as an alternative to current technologies in economically accessing ore bodies in excess of 3 500m.

Keywords : linear synchronous motors, ropeless hoists, ultra deep level gold mining, transport

1. INTRODUCTION

The use of linear motors for transport applications is gaining popularity. They have been used successfully in high-speed electric trains developing incredibly high thrust forces. An example of this is the Japanese research MAGLEV (Magnetic Levitation) MLX01 train that has set a record for these types of vehicles of 550 km/h on 12 December 1997. This is achieved with no physical contact between the mover and the stationary sections of the system.

Research into linear motors, with a view to hoisting applications for ultra deep level gold mining was begun at the Department of Electrical Engineering in 1994 with the successful design, construction and testing of a horizontal linear induction motor (LIM). This was not seen as a viable option primarily due to the non-linear control characteristics of the motor, low efficiencies and poor power factor. A long-primary, short secondary linear synchronous motor (LSM) was found to be the best configuration for hoisting applications, and a scaled prototype was constructed.

Research has centered around higher harmonic effects, buried- versus surface-mounted magnets on the secondary, reduction of Detent forces using various motor configurations and control strategies. All these studies relied heavily on finite element work (FEM) and measurement of the static thrust. The level of research can be gauged by the force-to-weight ratios (FWR) achieved by the various motors. The FWR is defined as the ratio of the force generated by the motor on the secondary to the actual weight of the secondary. Figure 1 gives an overview of the research conducted to date.

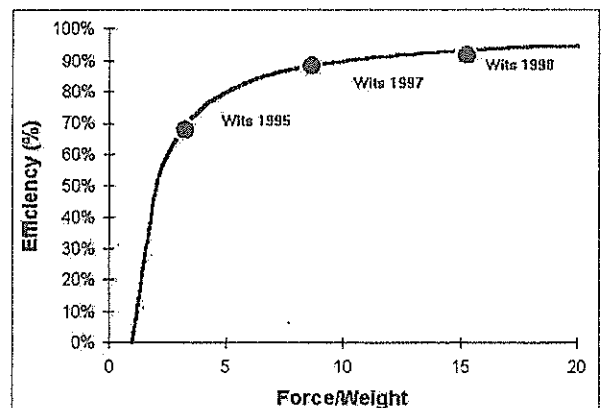


Figure 1 Improvement of FWR and efficiency of a LSM hoist

2. BACKGROUND

The gold resources of the Witwatersrand basin (South Africa's largest gold-ore body) are continuously depleting. For example in 1970 some 1 266 t of gold (79% of the world's production) was produced from this depository [1]. In 1992 production had decreased to 614 t of gold, and some 200-300 t are forecast for the year 2025 [2]. It is due to this depletion in production that the mining houses are being forced to consider resources at ultra-depth.

Depending on the rock strength, ultra-deep mining is defined as mining at depths exceeding 3 500m. It is possible to hoist economic tonnages from 4 000m depth in a single shaft system [3] and the auriferous orebodies, colloquially termed reefs, are known to extend to at least 6 000m below the surface[1]. Some mines are already mining at depths exceeding 3 500m. Access to these orebodies is achieved through multiple vertical shaft systems and horizontal haulage systems. Limiting factors in the horizontal plane are negligible

due primarily to the fact that it is based on proven technologies such as locomotives and diesel power.

However there exist major problems in the vertical plane. The hoisting system most commonly used has a large electrical drive mounted on the surface (usually a dc motor) which rotates the winding drum to which the lift cage is attached via the winding rope. Although there are many other problems with this type of system, the main limiting factor is the winding rope. To meet the stringent safety specifications of the South African Minerals and Regulation Act, the rope strength and hence the diameter has to be relatively large when compared to the weight of the payload. Thus after a certain depth, the hoisting rope can no longer support its own weight together with the payload it is meant to carry. Thus, the maximum depth achievable by a single hoist system is limited; currently to approximately 2 800 m (Figure 2).

To overcome this problem, sub-vertical shafts are sunk. A sub-vertical shaft uses exactly the same hoisting principles, with the only exception being that the electric drive is situated underground at a depth of around 2 000 m. This results in an increased capital outlay as well as an increase in the general operating costs due to the bottleneck, which occurs between the two shafts.

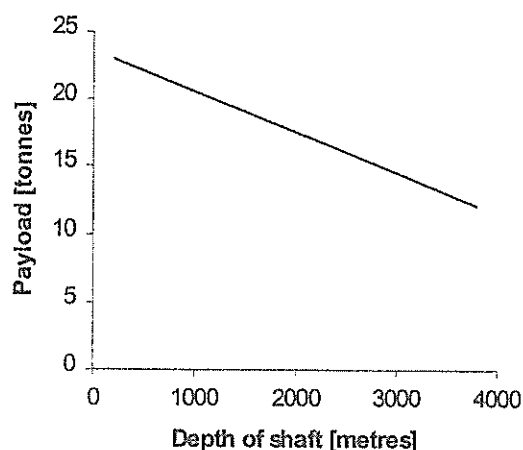


Figure 2 Effective payload of conveyance versus depth of compound strand winding rope for Blair multi-rope winder

3. UNDERGROUND TRANSPORT REQUIREMENTS

The transport infrastructure must be designed to cater for a number of commodities, not all of which necessarily require the same speed or mode of

transportation and are for different destinations (Table 1).

Table 1 Commodities to be transported in a mine

COMMODITY	DESTINATION	SPEED
Personnel	Surface to workplace	$15\text{ms}^{-1} - 20\text{ms}^{-1}$
	Workplace to surface	
Mining equipment/supplies	Surface to workplace	$15\text{ms}^{-1} - 20\text{ms}^{-1}$
	Workplace to surface	
Reef / Waste rock	Face to surface	$5\text{ms}^{-1} - 10\text{ms}^{-1}$

Currently, mining takes place simultaneously over a wide area of the mine, utilizing a number of levels and mining areas on each level. The scattered nature of mining requires a major underground network, generally tracked. This is usually grossly underutilized, and often poorly maintained. This system has to cater for the transport of personnel, mining equipment/supplies, support equipment and reef/waste rock.

In today's mines, face-workers have to travel, generally in two shafts. Some mines do have tertiary shafts. Depending on the layout, the sub-vertical shafts may be some distance from the previous shaft and the move from one shaft to the next is generally on foot. The consequence of this method of getting workers to their place of work is that up to three and a half hours per shift are lost in transportation.

The timeous transport of mining equipment to the many working areas where tracked transport underground is the primary/sole system available is made difficult by the number of operating levels, the need to transport reef and waste rock in the opposite direction of materials and supplies, the poor quality of track and tracked equipment maintenance and logistical difficulties in ensuring correct delivery of equipment and supplies. As a result of this, techniques such as *Just in Time* become increasingly more difficult to implement, together with the natural rise in the costs of inventory and loss of equipment.

Due to the scattered nature of mining practices, the quantities of reef and waste trammed from any specific area to the shaft is relatively small but does require significant time and effort to achieve. This is as a result of the long tramming distances, the relatively small payload per locomotive and the high level of shunting [4].

4. LINEAR SYNCHRONOUS MOTORS

Research has shown that the best configuration for a linear synchronous hoist is a long primary design with permanent magnet excitation on the lift cage [6]. The design consists of two long primary sections mounted on the side of the shaft and two short secondary

sections mounted on the side of the lift cage (Figure 3). The primary consists of copper windings mounted into modular steel sections. The secondary consists of two arrays of rare-earth permanent magnets (Nd-Fe-B35).

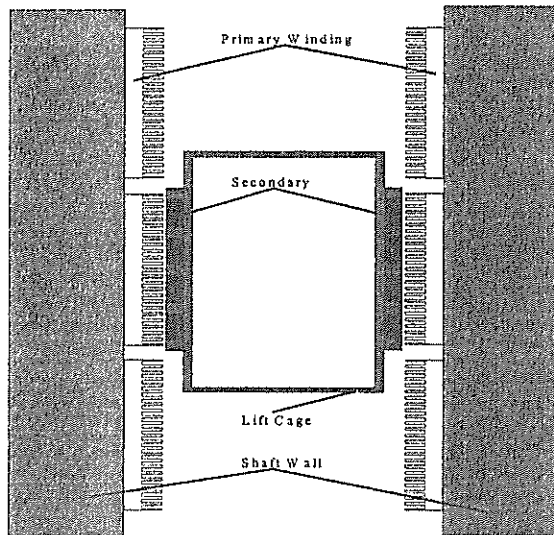


Figure 3 LSM in operation in a shaft

As the direction of travel is determined solely by the construction of the guide system, both horizontal and vertical transport is possible (Figure 4 b). It is envisaged that in mines of the future the size of conveyances is reduced dramatically to the extent that they become almost "pod"-like, capable of carrying one person or equivalent size in material. It is also proposed that the same conveyances be used for the transportation of material/supplies and rock, thus ensuring minimal empty conveyances on any one journey

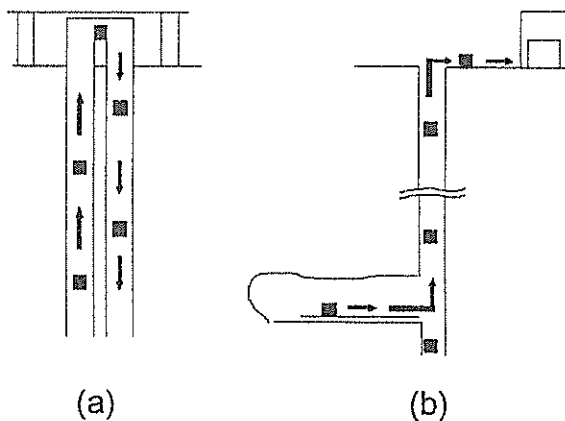


Figure 4 Linear motor in operation in a mine

5 ADVANTAGES

With the use of linear synchronous motors, underground transportation costs can be minimised and safety can be improved. To ensure minimal

energy consumption, only those windings which are in close proximity to the lift cage are excited, while the windings of the other cores are kept short-circuited. This ensures that if a power failure or a fault occurs, the dynamic braking characteristics of the system will significantly reduce the speed of the falling lift cage, providing a safe decent for both men and ore [5].

As opposed to conventional hoisting techniques, where the mass of the rope limits single shaft operation to approximately 3 500m, the proposed LSM hoisting is independent of depth, primarily due to the absence of a rope. However, as one goes deeper, considerations such as rock mechanics (to provide adequate support) and ventilation (to reduce virgin rock temperatures) needs to be addressed.

With the elimination of the steel wire rope, it is now possible in LSM hoisting for multiple conveyances to occupy the same compartment in the shaft area. Thus the cross sectional area of the mine shaft can be dramatically reduced and still satisfy the requirements of transportation of men/material and ore. Consequently different sinking techniques can be employed, such as raise boring, to reach these depths. This results in significant savings in terms of capital investment.

Owing to the fact that the direction of travel is dependent solely on the guide system, transportation in any plane (vertical, horizontal or incline) is possible. One benefits would be allowing access haulages and shafts to "follow the reef". This results in significant savings in development costs and travelling times.

Transportation of men is more continuous and is more comfortable and less physically demanding. Since the need to change conveyances or mode of transport is eliminated, transport times from surface to place of work and back to surface are reduced dramatically. As a result, the efficiency and productivity of the work force is improved substantially.

Transportation of equipment and supplies now lends itself to the full implementation of *Just in Time* and *Just in Place* principles. Thus, transportation of material can take place from surface store to underground destination in a minimum time. Further, the correct destination can be improved with an associated reduction in stock loss during transportation.

The transportation of reef to surface can now be achieved with a dedicated system. With tonnage requirements of a shaft in mind, the movement toward a batch process allows for a reduction in shaft diameter and individual conveyance speed. The

requirement for large storage facilities. is now eliminated, and separate transportation of waste rock and reef can now be ensured.

One of the problems associated with conventional hoisting systems, is the dedicated usage of a particular winder. Thus a rock winder will only convey ore from shaft bottom to surface, and a material winder will carry equipment and supplies from surface to underground. Although in a double drum system an empty conveyance moving in one direction is used as a counter balance for the full conveyance moving in the opposite direction. This results in the winder essentially only doing work on the load. However there are associated losses such as friction of the rope on the drum and sheave, gearbox losses, etc. With the elimination of the rope, and associated attachments, conveyances can now become multi-purpose vehicles, with the same unit transporting men, material and ore. This design allows for an improved equalization of mass thus reducing the requirements of empty conveyances on any one trip.

Due to the fact that the speed of a synchronous motor is a function of the fundamental supply frequency, transportation can be accomplished in a rapid manner. Thus, whether the LSM is driven from an inverter bank or a rotary synchronous generator, dynamic control of the transportation fleet can be achieved.

In conclusion, the reduction in size of conveyance lends itself to the improvement of organizational control over the entire 'rolling stock' and allows for the incorporation of information technology into management techniques [7].

6. CONCLUSION

Having successfully completed testing on various small scale models, the next step is to design and construct a full-scale model. System and element optimisation can then be conducted.

The linear synchronous motor has many advantages not only in terms of economic considerations, but also with respect to safety. One of the more attractive features is the regenerative braking characteristic of

the motor, making it highly attractive for one of the most arduous environments known to man. Linear motors in general and in particular synchronous motors offer a viable alternative to current hoisting technologies on South African gold mines in accessing orebodies in excess of 3 500m.

Although the linear synchronous motor has the technological ability to reach these depths, additional research has to be completed before it becomes a reality. Environmental issues need to be addressed where virgin rock temperatures are in excess of 50° C, as well as rock mechanic considerations where movement of the earth at these depths becomes a major concern [4].

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COMPARISON OF RELUCTANCE SYNCHRONOUS AND INDUCTION MOTOR SUBURBAN LOCOMOTIVE DRIVES

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Abstract: The paper considers the preliminary results of an investigation into the possible use of a reluctance synchronous machine as a traction motor for suburban locomotives. The intention is to compare the performance of a 110 kW induction locomotive traction motor with that of a reluctance synchronous traction motor. The same stator is used for both motors. The rotor of the reluctance synchronous motor is optimum designed with 2-D finite element software and optimisation algorithms. The paper give some calculated performance results of the reluctance synchronous motor in comparison with the induction motor.

Keywords: RSM = Reluctance synchronous machine, IM = Induction machine

1 INTRODUCTION

There is a renewed interest in the design of Reluctance Synchronous Motors (RSM) aimed at obtaining the same or better performance than its Induction Motor (IM) counterpart. Despite the lower power factor and probable higher kVA inverter rating of the RSM drive, the RSM drive has attractive advantages from a machine point of view including, amongst other things, a good torque density and efficiency, a very simple current vector controller and low manufacturing costs. A comparative study of the performance of the RSM and the IM in the low power range [1] has shown that the torque of the normal laminated rotor RSM is superior to that of the IM. Recently it has been shown [2] that in the medium power range (42 kW) the torque density of the inverter-fed RSM is more than 30% higher than that of the standard, sinusoidal-fed IM for the same amount of total copper losses. Keeping in mind that the comparison of [2] is not fair, because a double-layer stator winding RSM is compared with a single-layer stator winding IM and furthermore the performance of an inverter-fed RSM is compared with that of a sinusoidal-fed IM, the torque improvement of the RSM will be even better.

There is a need for a true comparison between the performances of inverter-fed RSMs and inverter-fed IMs. It is well-known that inverter-fed induction machines are in general derated due to, amongst other things, additional harmonic copper losses in the rotor [3]. These additional harmonic rotor losses are not present in the inverter-fed RSM, which is a very important advantage. The comparison between the RSM and IM must also be directed at higher power levels, such as for traction applications. This is the aim of an investigation being conducted at the

University of Stellenbosch into the performance capability of inverter-fed RSMs and IMs for suburban locomotives.

The investigation is at a rated power level of 110 kW. An optimum designed 110 kW IM traction motor for suburban locomotives is obtained for this investigation. For the first comparison the same stator, namely the stator of the 110 kW IM, is used for both the IM and the RSM. For the RSM the rotor of the IM is removed and replaced by an optimum designed, normal laminated flux barrier rotor. This paper describes the optimum design of the RSM rotor and gives the performance results obtained in comparison with the IM.

2 INDUCTION TRACTION MOTOR

The specifications of the IM traction motor are given in Table 1. Shown in Fig. 1 is the cross-section of the stator lamination. The stator winding is a double-layer winding of which the windings are chorded with a 8/9 pitch. The stator has 54 slots with 3 slots per pole per phase. It has an outer diameter of 485 mm, an inner diameter of 300 mm and the stack length of the stator is 340 mm. The air holes in the stator lamination are also shown in Fig. 1. The purpose of the holes is to circulate the air-flow within the machine using an internal shaft-mounted cooling fan.

Table 1: Specifications of the IM.

Machine specification	Rated value
Number of poles	6
Supply voltage	500 V
Supply current	204 A
Torque	702 Nm
Output Power	110 kW
Speed	1496 r/min

3 RELUCTANCE SYNCHRONOUS TRACTION MOTOR

As mentioned above the stator of the IM (Fig. 1) is also used for the RSM traction motor. A cross-section

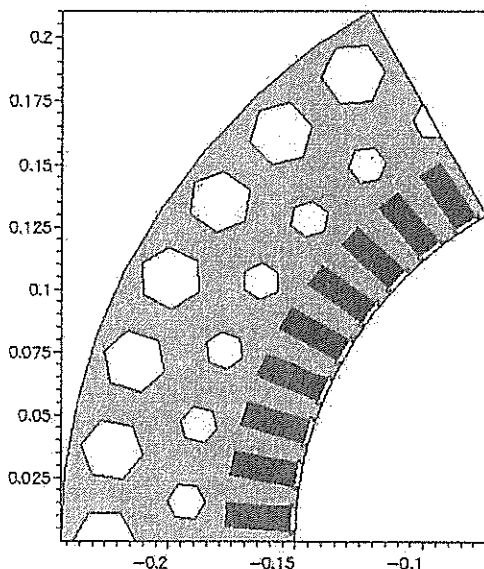


Figure 1 One pole segment of stator.

of the rotor of the RSM is shown in Fig. 2. It consists of four airgap slots, so-called flux barriers, to create magnetic asymmetry on the rotor. The parameters defining the shape of the flux barriers are the barrier height (bh), the web width (ww), the rib height (rh) and barrier width (bw). The height of the barriers, bh , is measured along the q-axis of the rotor from the centre of the shaft. The barrier height, bh , and the barrier width, bw , are the variables of the rotor to be optimised. The positions of the flux barriers at the surface of the rotor are fixed (described further in section 4).

The torque equation of the RSM is given by

$$T = K\Delta L(\phi) I_s^2 \sin(2\phi) \quad (1)$$

where ΔL is the inductance difference, $L_d - L_q$, I_s is the rms stator current and ϕ is the current angle between the current space phasor and the d-axis of the rotor in the dq reference frame. K is a machine constant. Note that ΔL is some function of ϕ [4]. The torque of the RSM is thus a function of the current angle, which is a third variable to be optimised to maximise the torque of the machine.

4 OPTIMUM DESIGN OF RSM ROTOR

There is a limit to the number of flux barriers per pole, N_{bp} , that can be used. To ensure that iron losses do almost not occur at all in the iron segments, a general guideline is to use eqn (2). N_{sl} is the number of stator

slots and p the number of poles. For a six-pole, 54-slot stator the number of barriers per pole is chosen as $N_{bp} = 8$, which is shown in Fig. 2.

$$N_{bp} < \frac{N_{sl}}{p} \quad (2)$$

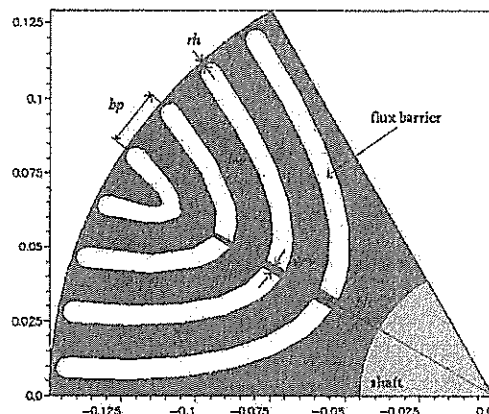


Figure 2 RSM rotor geometry.

With the number of flux barriers per pole known and with an equal distribution of the flux barriers in the rotor, the positions of the flux barriers at the surface of the rotor are known. Hence, the two positions or points of a flux barrier at the rotor surface are known and fixed. A third position or point of the flux barrier is chosen on the q-axis of the rotor. This is given by the distance bh in Fig. 2 and is a variable to be optimised. A parabola is fitted through the three points to give the shape of the flux barrier.

4.1 Mechanical strength optimisation

The rotor rib height (rh) and web width (ww) are optimised manually with minimum dimensions as objective function during mechanical finite element stress analyses. In these analyses the maximum Von Mises stress calculated is constrained to be a factor 2 lower than the yield strength of the material used for the rotor lamination. The analyses showed that the web of the top flux barrier is not necessary. The stresses are calculated at a maximum speed of 4500 r/min.

4.2 Torque optimisation

The objective function value to be maximised is the torque per copper losses of the RSM. The torque is directly calculated in the finite element software program. The copper losses are taken as the stator copper losses at rated stator current (204 A) and peak stator current (350 A). The variables to be optimised are already mentioned as the barrier width (bw), the barrier height (bh) and the current angle (ϕ). The

algorithm of Powell is used in the optimisation procedure.

The finite element program, which calculates the output function values of torque, efficiency, power factor, etc., together with the optimisation algorithms used, are described in detail by [5]. The software is run on a 233 MHz Pentium II computer with 64 Mbytes RAM. The finite element software, which is not of commercially available, makes use of triangular elements of the first order. Only one pole of the motor is meshed with one airgap macro-element comprising nodals on both sides of the airgap. A time-saving scheme makes use of one airgap element to model rotor movement. The Newton Raphson method is used for the solution of the set of nonlinear equations.

5 CALCULATED RESULTS

The results from the optimisation using Powell's method showed that the top flux barrier can be removed, resulting in the same performance. This means that only three flux barriers on the position of four are used in the rotor. The final geometry of the rotor lamination is shown in Fig. 3. Fig. 4 shows a typical finite element field solution obtained with the software. The initial and final values of the variables optimised are given in Table 2. The design of the RSM is optimised for two cases, namely the maximisation of the torque per *rated* stator copper losses (at 204 A) and the maximisation of the torque per *peak* stator copper losses (at 350 A).

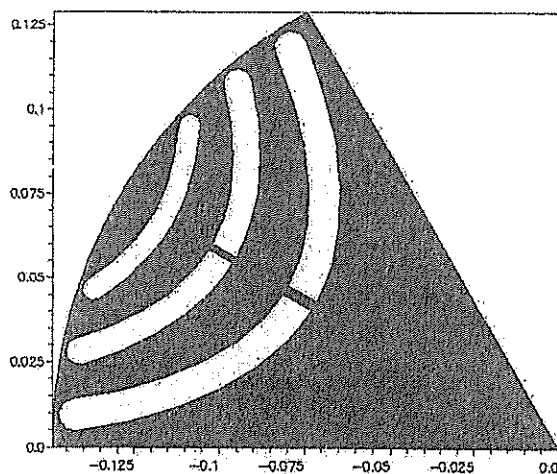


Figure 3 Optimised geometry of RSM rotor.

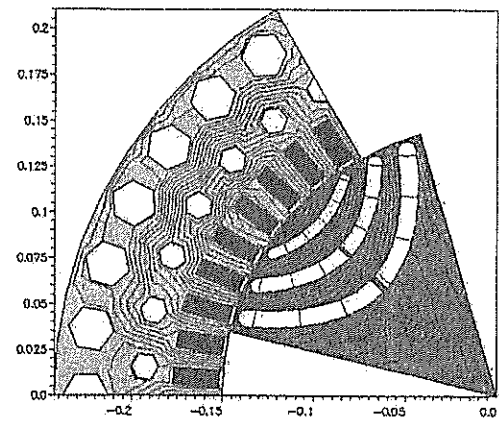


Figure 4 Finite element field solution.

From the results of Table 2 it is clear that there is little difference between the optimum dimensions for the two cases. The reason for this lies in that a relative large quantity of iron is used on the stator and the rotor so that there is little difference in the magnetic saturation level of the steel when using rated or peak current. The upper barrier width in the case of peak current optimisation (with peak copper losses) is slightly wider, probably to reduce the higher cross magnetisation. The barrier heights are slightly less to obtain more iron on the rotor for less magnetic saturation. The optimum current angle increases in the case of peak current optimisation, to reduce the magnetic saturation (a higher current angle implies less d-axis current or else less magnetisation current).

Table 2: Optimised variables.

Variable [mm]	Start	$T_{I=204A}$	$T_{I=350A}$
$bh(1)$	65.00	89.21	88.14
$bh(2)$	85.00	114.72	114.31
$bh(3)$	105.00	135.06	134.50
$bh(4)$	125.00	-	-
$bw(1)$	7.20	9.56	9.50
$bw(2)$	7.20	8.04	8.04
$bw(3)$	7.20	5.88	6.36
$ww(1)$	1.80	1.50	1.50
$ww(2)$	1.50	1.25	1.25
$ww(3)$	1.25	1.25	1.25
$ww(4)$	1.00	-	-
ϕ	60.00°	59.43°	64.00°

In Fig. 5 the output torque of the RSM as a function of the current angle is shown in the cases of rated and peak current in the machine. With the machine operated with peak current the optimum current angle increases from 59° to 64°. The current angle increases to avoid deep magnetic saturation of the steel. A maximum torque of 670 Nm at rated current and a maximum torque of 1345 Nm at peak current are obtained with the RSM traction motor. This is to be compared with the torque outputs of the IM traction

motor with the same stator currents, namely 702 Nm and 1500 Nm respectively. These results are compared in Table 3.

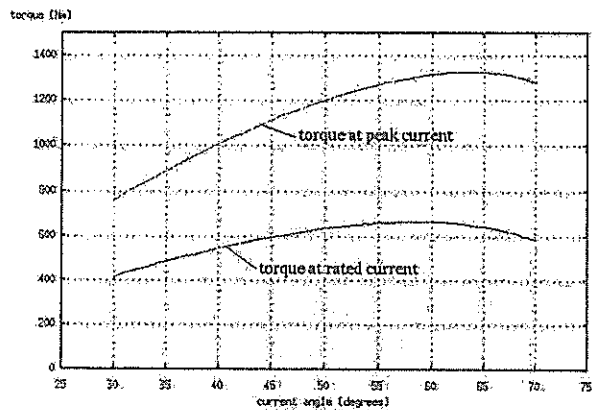


Figure 5 Torque at rated and peak current.

Table 3: Torque at rated and peak current.

Current [A]	RSM [Nm]	IM [Nm]
204	670	702
350	1345	1500

The RSM has no rotor copper loss and therefore the input current of the RSM, compared to the IM, can be

raised by a factor, $\sqrt{\frac{P_{\text{total copper loss(IM)}}}{P_{\text{rotor copper loss(IM)}}}}$, to obtain

the same total copper losses and roughly the same operating temperature (the latter has to be confirmed by measurements). It is shown in Fig. 6 that in the RSM with a constant, but close to optimum current angle, there is almost a linear relationship between torque and *rms* current. This was also found by [2]. If the current of the RSM, thus, can be raised, the output torque will also rise similarly. This then will result in a higher output torque than that of the IM.

Furthermore, the use of an IM stator for the RSM is to the detriment of the performance of the RSM [1]. If an optimum designed RSM stator is used, then the torque output performance of the RSM will improve even further.

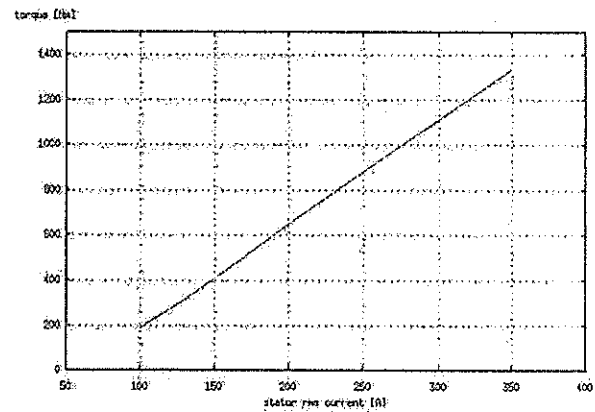


Figure 6 Torque vs *rms* stator current.

Fig. 7 shows the optimised result of the mechanical finite element strength analysis of the rotor lamination. The deformation result shows that the maximum radial displacement of the lamination at a speed of 4500 r/min is 80 μm , which is safe for a pre-chosen airgap of 0.9 mm.

6 CONCLUSIONS

The preliminary comparison results showed that without optimising the stator (using an IM stator), the RSM's performance compares well with that of the IM. The latter is so despite the fact that using an IM stator for the RSM is to the detriment of the performance of the RSM.

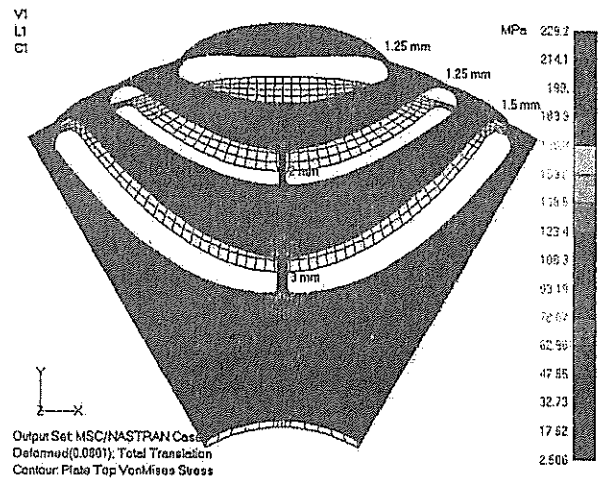


Figure 7 Mechanical finite element stress analysis displacements (blown up)

As far as the cooling of the RSM is concerned, the rotor flux barrier openings can be used to direct the airflow in the machine in a similar way as in the case of the IM. The advantage of the flux barrier cooling openings is that part of these openings is close to the stator where all the heat is generated. The cooling, thus, of the RSM is expected to be more efficient.

Power and temperature measurements have to be taken to compare the performance of the inverter-fed RSM with that of the inverter-fed IM. By doing so the effect of harmonic-induced rotor currents in the IM due to the inverter supply can be taken into account in the comparison.

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AN ARTIFICIAL NEURAL NETWORK IDENTIFIER FOR THE CONTROL OF TURBOGENERATORS

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Abstract: The increasing complexity of modern power systems highlights the need for advanced modelling techniques for effective control of power systems. This paper presents results of simulation studies carried out on identifying the dynamics of a single turbogenerator connected to an infinite bus through a short transmission line, using a Continually Online Trained (COT) Artificial Neural Network (ANN). The emphasis in this paper is on the type of ANN identifier inputs and outputs that are required and found sufficient for an ANN control.

Keywords: Multilayer perceptron network; artificial neural network modelling; power plant control.

1 INTRODUCTION

Synchronous turbogenerators supply most of the electrical energy produced by mankind and are largely responsible for maintaining the stability and security of the electrical network. The effective control of these devices is, therefore, important. However, a turbogenerator is a highly non-linear, fast acting, multiple input multiple output device with a wide range of operating conditions and dynamic characteristics that depend on the power system to which the generator is connected [1,2]. Conventional automatic voltage regulators and turbine governors are designed to control, in some optimal fashion, the turbogenerator around one operating point; at any other point the generator's performance is degraded.

Adaptive controllers for turbogenerators are usually designed using linear models and traditional techniques of identification, analysis, and synthesis to achieve the desired performance. Often restrictive assumptions are made [3] about the disturbance that the system is likely to be subjected to. However, due to a turbogenerator's wide operating range, its complex dynamics, its transient performance, and its non-linearities, it cannot be accurately modelled as a linear device. This paper explains how a COT ANN can be used to identify the turbogenerator from moment to moment to enable effective control at all operating points. Results of simulation studies using two different sets of input signals for the ANN identifier (IANN) are presented in this paper. This paper reports that for the second set of IANN input signals the Multilayer Perceptron (MLP) network can also track, albeit with reduced accuracy, the terminal voltage and speed deviations of the turbogenerator even when the training is terminated unlike with the first set of IANN input signals where no tracking is seen.

2 LABORATORY SYSTEM BEING TESTED

The ANN identifier is tested on a simulated model of a 3 kW Mawdsley micro-alternator [4] shown in Fig. 1. The micro-alternator has per-unit parameters which are typical of those normally expected of 30 - 1000 MW generators, and it is also equipped with a dc motor prime mover controlled to act as a turbine simulator.

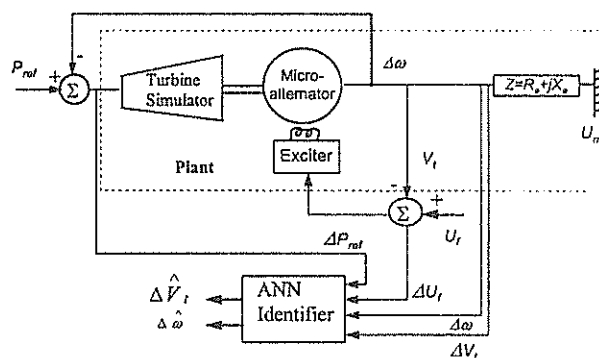


Fig. 1 Laboratory system of a single machine connected to an infinite bus through a short transmission line.

The micro-alternator system in Fig. 1 is without any automatic voltage regulator or governor controllers, and can be described by the dq-axis equations with the machine currents, speed and rotor angle as the state variables; it has one damper winding on each axis thus giving a seventh order model. The exciter is modelled as a first order device with limits on its output voltage levels. The turbine (simulator) is also modelled as a third order device which includes reheating between the high pressure and intermediate pressure stages. The output of the turbine simulator is limited between zero and 120%. A relatively short transmission line

connecting the generator to the infinite bus is modelled using the eqs. (1) and (2) in the state space form.

The micro-alternator, exciter and turbine form the eleventh order turbogenerator which is the so-called *plant* to be identified in this paper using a COT ANN. In the simulation process these eleven first order differential equations are solved using MATLAB/SIMULINK.

$$u_d = U_m \sin \delta - R_e i_d - X_e i_q - L_e i_d \quad \dots(1)$$

$$u_q = U_m \cos \delta - R_e i_q + X_e i_d - L_e i_q \quad \dots(2)$$

where u_d and u_q are voltage components at the machine terminals, U_m is the peak value of the sinewave voltage at the infinite bus and R_e , X_e , L_e are transmission line parameters.

3 COT ANN PLANT IDENTIFIER

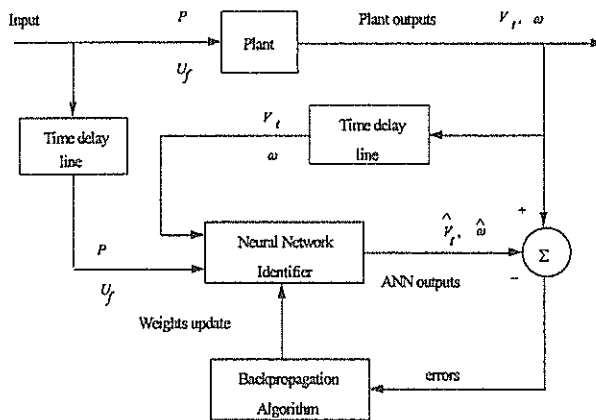


Fig. 2 COT ANN Identifier

The Identifier ANN (IANN) in Figs. 2 and 3 is a feedforward MLP network and has three layers consisting of an input layer with twelve inputs, a single hidden layer with sigmoidal activation functions consisting of fourteen neurons and an output layer with two outputs. The first set of inputs to the IANN are also the *actual* power input signal P to the turbine simulator, the *actual* field voltage input U_f to the exciter, the *actual* terminal voltage V_t and the *actual* speed deviation $\Delta\omega$ of the generator as shown in Fig.2. These four ANN inputs are delayed by the sample period of 20 ms and, together with eight previously delayed values, form twelve inputs altogether to the IANN. A sampling frequency of 50 Hz is chosen which is sufficiently fast for the IANN to reconstruct the speed and terminal voltage signals from the sampled input signals since the natural

oscillation frequency of the turbogenerator speed deviation is about 3 Hz and the response of the turbogenerator to the terminal voltage changes is even slower and is about 0.3 Hz. For the first set of IANN inputs above the IANN outputs are the *estimated* terminal voltage $\hat{V}_t$ and *estimated* speed deviation $\Delta\hat{\omega}$ of the generator.

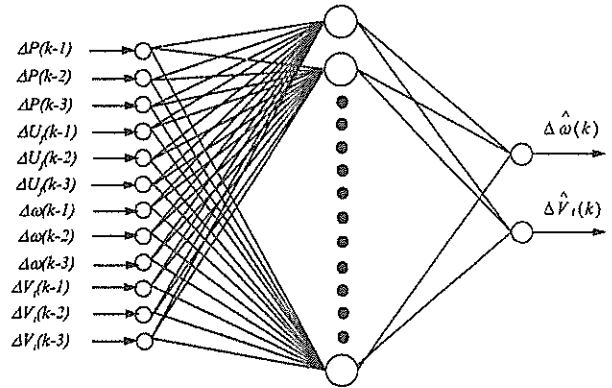


Fig. 3 IANN architecture

The second set of IANN inputs are also the *actual* deviation ΔP in the input to the turbine simulator, the *actual* deviation ΔU_f in the input to the exciter, the *actual* terminal voltage deviation ΔV_t and the *actual* speed deviation $\Delta\omega$ of the generator. These four ANN inputs are also delayed by the sample period of 20 ms and, together with eight previously delayed values, form twelve inputs altogether to the IANN as shown in Fig. 3. For the second set of IANN inputs above the IANN outputs are the *estimated* terminal voltage deviation $\Delta\hat{V}_t$ and *estimated* speed deviation $\Delta\hat{\omega}$ of the generator.

The number of neurons in the hidden layer of the IANN was determined empirically. The IANN weights were set to small random values and the conventional backpropagation algorithm was used to update these weights of the IANN. The differences between the respective *actual* outputs of the turbogenerator model and the *estimated* outputs of IANN form the error signals for the updating of weights in the IANN. A reasonable learning rate was determined by training this neural network and setting the learning rate parameter so that a compromise between the training time and the accuracy of the network is achieved.

4 SIMULATION RESULTS

The training of the IANN was simulated using pseudorandom binary signals generated in MATLAB

and fed into the exciter and the turbine. These random signals excite the full range of the dynamic response of the turbogenerator. A constant field voltage U_f and a turbine power signal P are applied to the plant, and disturbances in the field voltage ΔU_f and in the turbine power signal ΔP are applied for training the IANN. The training signal ΔU_f applied to the exciter is shown in Fig. 4. The errors after 20 s of training are insignificantly small. The initial weights for the IANN were set to some random values in the range of $[-0.1, 0.1]$ to achieve fast learning of the plant dynamics. A learning gain of 0.05 was used for the backpropagation algorithm. The IANN is only required to generalise one time step ahead, that is 20 ms, and therefore no momentum term was used.

The tracking capabilities of the IANN were tried out by terminating the backpropagation training after 25 s, but continuing with the simulations of the turbogenerator model and the IANN for a further 5 s and then switching back to *online* training mode. Figs. 5 and 6 show that for the first set of IANN inputs the neural network is unable to track the desired outputs when the training is terminated. Fig. 7 is enlarged graph of Fig. 6 showing the non-tracking capability by the IANN of the actual speed deviation between 20 s and 30 s when the *online* training is terminated. This means that the training of IANN has led to an extreme *adaptive local minimum* which is dangerous if the ANN training has to stop for some unforeseen circumstances.

Figs. 8 and 9 show that for the second set of IANN inputs the neural network is able to track the outputs even when the training is terminated. Fig. 10 is enlarged graph of Fig. 9 showing that the IANN is able to track, albeit with reduced accuracy, the speed deviation likewise the terminal voltage deviation even when the training is terminated between 20 s and 30 s. This is as a result of using deviations as inputs to the ANN identifier. This means that the training of IANN with deviation signals has led to some extent an *adaptive global minimum*. This is found more suitable for implementing the ANN controller in [5] because *online* training of the ANNs will only be carried out when the operating point of the turbogenerator changes and therefore ensuring minimum controller drift which otherwise can lead to instability of power system.

5 CONCLUSIONS

Early conclusions of this work indicate that the COT ANN can model the turbogenerator dynamics when the network configuration and system operating point changes. The successful identification of the turbogenerator dynamics by the COT ANN occurs

because *deviation signals are used as inputs to the IANN* and the *online training never stops*. The COT ANN identifier can be used in conjunction with a separate neural network controller to allow greater usage of existing power plant by effective control of the excitation voltage and turbine power of a turbogenerator [5].

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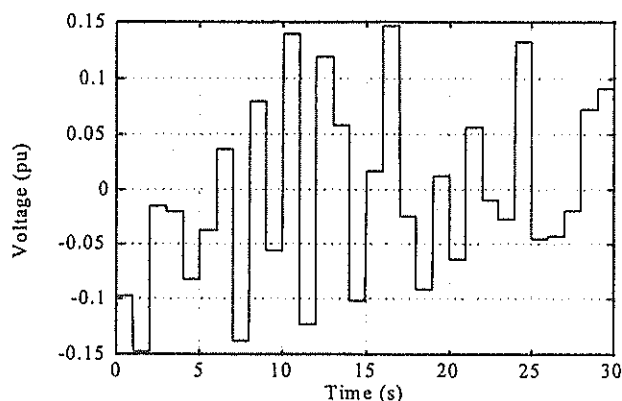


Fig. 4 Training signal applied to the exciter

FLICKER EMISSION VARIATIONS AT AN OPEN-ARC SCRAP MELTING FURNACE

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Abstract

This paper investigates variations in the characteristic flicker emission coefficient, K_{st} , and the flicker severity, P_{st} , of an electric arc furnace plant. The study is based on flicker measurements conducted at a 60 MVA AC scrap melting open-arc furnace. The impact of a series reactor installed at the furnace is determined and the correlation between the flicker emission coefficient and the charge material size loaded into the furnace is evaluated. Additionally, the influence of the amount of charge material in a basket on the flicker emission per basket is investigated and a modified flicker prediction model is suggested for use at arc furnace plants.

Keywords:

Flicker, Flicker emission coefficient, K_{st} , Arc furnace.

1 Introduction

Flicker refers to a visual sensation experienced by humans subjected to varying illumination intensity. This variation in luminance is induced by a varying supply voltage connected to electrical sources of light energy, such as a 60 W tungsten coiled, gas-filled, filament lamp supplied by a 230 V supply voltage. The discernment of flicker by the human visual system is dependent on the magnitude of illumination intensity changes, the frequency at which these illumination intensity changes occur and the human visual system that discerns these illumination changes. [1]

The voltage of the supply network is controlled at the generation point to a specific nominal value. The supply authority does not operate any equipment within the power system that produces voltage fluctuations in the flicker frequency range of 0,5 Hz to 35 Hz. The only possible cause of repetitive voltage fluctuations is found at the industrial consumer [1].

An electrical load connected to the supply network will require current to flow from the generation station to the load. Figure 1 illustrates the voltage drops induced by the load current over transformers, transmission lines and any other electrical supply equipment. Changing the tapping points on the supply transformers and connecting capacitors onto the network can, under normal circumstances, compensate for the voltage drops in the supply network.

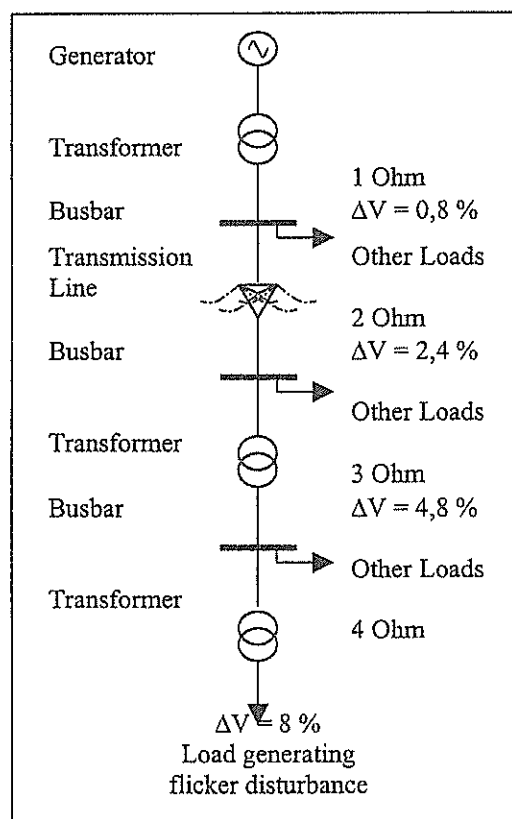


Figure 1: Distribution of flicker in the supply network

When a varying load is connected to the supply network, the load will generate flicker on the supply network if the load fluctuations are of a significant magnitude and occur at a frequency within the flicker frequency range. Transformer-taps, which are normally used to compensate for voltage drops in the supply network, cannot be changed fast enough to compensate for voltage fluctuations in the flicker frequency range. The flicker caused by the fluctuating load will consequently not be compensated by the transformer taps and will be propagated through the supply network [1]. Switching capacitors in and out of the supply network will also not be useful in eliminating flicker.

2 FLICKER FROM ELECTRIC ARC FURNACES

Flicker is normally thought of as a problem associated with the quality of supply from a utility. It is however playing an ever-increasing role in the design and operation of electric arc furnace plants [2]. Arc furnaces produce significant amounts of flicker and are the most critical loads as far as flicker emission is concerned [1], [3].

An arc furnace uses electric arcs to convert electrical energy to heat energy. The electric arc is an electrical discharge in a gas, characterized by a low voltage and a high current, with a non-linear relationship between voltage and current. All current fluctuations within the arc, will cause voltage fluctuations in the network and will result in the emission of flicker from the arc-furnace. An open-arc scrap-smelting furnace, for example, produces load variation between the extremes of a brief full short circuit to an open circuit within milliseconds. This results in a high flicker emission level from the arc furnace.

3 THE FLICKER PREDICTION FORMULA

3.1 Derivation of the flicker prediction formula

To obtain an indication of possible flicker problems that may be encountered when a new arc furnace plant is developed, the relation between the electrical size of the proposed arc furnace and the fault level of the supply network may be used. The electrical size of the EAF can be expressed by its short-circuit power S_{ccf} , which is normally used, or by the maximum apparent power measured at an existing plant. The relation S_{ccf} / S_{ccn} is referred to as the short circuit voltage depression for the arc furnace [4].

Figure 2 shows a simple equivalent diagram for a supply network feeding an electric arc furnace. Figure 2(a) on the left represents an open circuit condition at the arc furnace where there is no current flowing to the arc furnace from the supply network. Figure 2(b) represents the current flow and voltage

drops through the circuit when there is a short circuit condition in the furnace, which allows maximum current to flow to the arc furnace. The change in current between the two conditions is equal to the current I_f flowing during the short circuit condition.

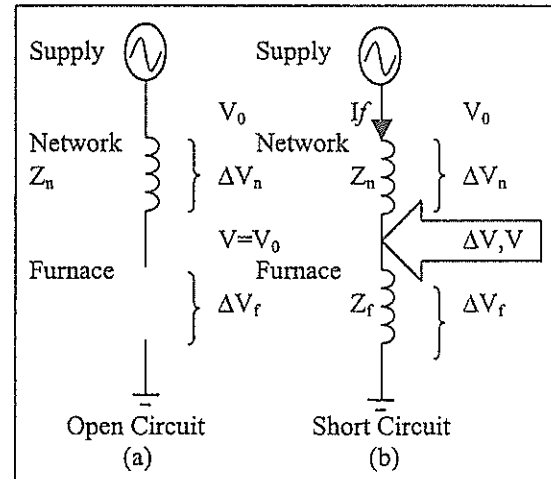


Figure 2: Simple equivalent supply diagram

To determine the flicker severity caused by the furnace, the magnitude of any voltage fluctuation is used and scaled as shown in the following equation:

$$P_{st} = f(\Delta V / V) \quad (1)$$

From figure 2 it is clear that $I_f = 0$ during an open circuit and that the voltage V at the furnace equals the supply voltage V_0

$$V = V_0 \quad \Delta V = \Delta V_n = 0 \quad I_f = 0 \quad (2)$$

From figure 2 it is also clear that the following apply to a short circuit:

$$V = V_0 - \Delta V \quad (3)$$

$$\Delta V = \Delta V_n = Z_n \cdot I_f \quad (4)$$

$$I_f = V_0 / (Z_n + Z_f) \quad (5)$$

For the voltage drop from zero load to load-on (short circuit) follows:

$$\Delta V / V_0 = Z_n / (Z_n + Z_f) \quad (6)$$

and for the voltage rise from load-on (short circuit) to zero load follows:

$$\Delta V / V = Z_n / Z_f \quad (7)$$

Z_n is comparatively small in relation to Z_f , therefore

$$\Delta V / V_0 \approx \Delta V / V \quad (8)$$

The short circuit power of the furnace (S_{ccf}) is given by:

$$S_{ccf} = \frac{V^2}{Z_f} \quad (9)$$

And the fault level of the network (S_{ccn}) as:

$$S_{ccn} = \frac{V^2}{Z_n} \quad (10)$$

The short circuit voltage depression (SCVD) of the arc furnace is defined as:

$$SCVD = \frac{S_{ccf}}{S_{ccn}} = \frac{V^2}{Z_f} \cdot \frac{Z_n}{V^2} = \frac{Z_n}{Z_f} \quad (11)$$

The short circuit deviation is the maximum voltage drop that can occur during furnace operation. Therefore, the flicker emission $P_{st99\%}$ of the furnace is assumed to be proportional to this value:

$$P_{st} = f(\Delta V/V) = K_{ST} \cdot S_{ccf} / S_{ccn} \quad (12)$$

where K_{ST} is a proportionality factor that depends on the special operational and constructional characteristics of the furnace.

To develop a method of predicting the flicker emission of an arc furnace, the SCVD of the furnace must be used while accounting for other possible flicker attenuation factors within the supply network. The most recent proposed model to determine the flicker severity level is given below [5]:

$$P_{st99\%} = K_{st} \cdot \frac{S_{ccf}}{S_{ccn}} \cdot \frac{1}{R_{comp}} \cdot C_{HV/LV} \quad (13)$$

where $P_{ST99\%}$ = Short term flicker severity not exceeded during 99 % of the measurement period.

K_{st} = Linear flicker emission coefficient

S_{ccf} = Arc furnace short circuit power

S_{ccn} = Short circuit power at the PCC

R_{comp} = Flicker reduction factor due to flicker compensation

$C_{HV/LV}$ = Flicker HV to LV transfer coefficient

3.2 Proposed modification of the flicker prediction formula

The flicker prediction model (13) assumes that the maximum load fluctuation that can occur at the arc furnace is when the furnace is subjected to a three-phase short-circuit followed by an open-circuit. S_{ccf} relate to the measurement result obtained while the three furnace electrodes were dipped into a molten bath to achieve a three-phase short circuit.

In the following study it is suggested that the actually measured maximum apparent power load demand of the furnace should be used, instead of the short circuit power of the furnace, to obtain an adjusted flicker emission coefficient for the furnace. This approach avoids the necessity of doing a dip test and takes all operational conditions automatically into account. The proposed flicker prediction model is shown in equation (14):

$$P_{st99\%} = K_{st} \cdot \frac{S_{max}}{S_{ccn}} \cdot \frac{1}{R_{comp}} \cdot C_{HV/LV} \quad (14)$$

where $P_{ST99\%}$ = Short term flicker severity not exceeded during 99 % of the measurement period.

$K_{st\ Adjust}$ = Adjusted flicker emission coefficient

S_{max} = Maximum apparent arc furnace power measured during the measuring period. The highest flicker peaks occurred when the first basket was loaded.

3.3 Calculation of S_{ccf}

The short circuit power of the furnace is quadratically related to the secondary voltage of the furnace transformer. Figure 3 shows the variation in the S-locus of the furnace when the secondary voltage of the transformer is increased and decreased:

$$S_{ccf} = V^2 / X \quad X = X_n' + X_f'$$

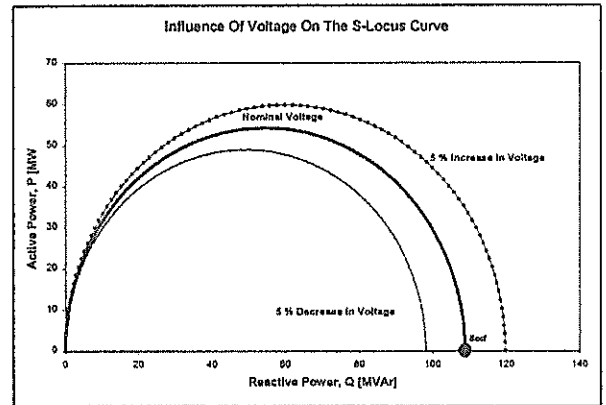


Figure 3: Variations in the S-locus of a furnace if the secondary voltage of the furnaces is adjusted.

S_{ccf} equals the circle diameter of the S-locus of the apparent power. The diameter changes with adjustments of V (e.g. due to different voltage taps) and / or with variations of X , which may for example be due to the introduction of a reactor in series with the furnace.

The apparent power measurements were adjusted to correlate with the nominal voltage of the measuring point. The apparent power measurements consequently represent the load when nominal voltage is supplied to the load and the actual voltage supplied to the load was not used.

3.4 Factors influencing the flicker emission coefficient

A good approximation of $K_{ST} = 75$ is suggested for the flicker prediction of the furnace before more detailed analyses can be conducted [5]. The flicker emission characteristics of an existing arc furnace plant are influenced by factors such as the availability of a series reactor and the charge material size melted in the furnace. Smaller and lighter charge material is more likely to be moved or tossed away by the arc, thus, causing arc length variations with higher instability and flicker emission than in the case of larger and heavier material.

4 The flicker prediction coefficient study

During November 1997 to February 1998 Davsteel afforded the Vaal Triangle Technikon an opportunity to investigate the impact of a new series reactor on the power quality of their 60 MVA arc furnace. The study, performed in collaboration with Rapha Pretorius and Associates, was designed to (i) investigate the flicker emission coefficient of the arc furnace, (ii) determine the impact of a new series reactor on the flicker emission coefficient and (iii) investigate relations between the flicker emission from the arc furnace and the charge material melted within the furnace. A flickermeter manufactured by the Technical University of Graz, and complying with IEC 868 specifications, was used to do measurements for the study [6].

The flicker reduction factor due to flicker compensation (R_{comp}) equaled one, because no flicker reduction equipment is installed at the arc furnace, and the flicker HV to LV transfer coefficient (CHV/LV) was assumed equal to one due to the closeness of Vanderbijlpark City.

4.1 Background information of the arc furnace

Davsteel uses a 60 MVA open arc scrap melting AC arc furnace to melt scrap, bales, cast iron and sponge iron. The liquid metal obtained from the arc furnace is refined in an 8 MVA ladle furnace to produce steel. The Western Vaal Metropolitan Substructure supplies electrical energy to the furnaces and the Davsteel point of common coupling (PCC) is at the 88 kV Vanderbijlpark "Munic" Substation, where the fault level is 3 800 MVA.

Flicker measurements combined with active and reactive power measurements were conducted at the

Davsteel 33 kV substation, where the fault level was 532 MVA. Neither the Western Vaal Metropolitan Substructure, nor Davsteel, was able to provide more than one voltage measuring point to enable the measurement of flicker emission from the furnace. The flicker level, measured at the Davsteel 33 kV substation, was consequently equated to the flicker emission level. The reticulation diagram for the 60 MVA furnace is shown in figure 4.

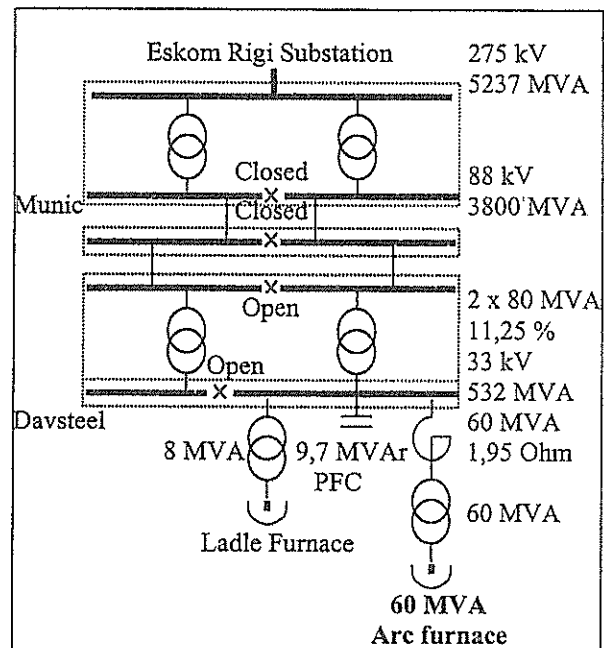


Figure 4: Reticulation diagram of the Davsteel 60 MVA arc furnace

A series reactor was installed at the arc furnace's 33 kV busbar in December 1997. The inclusion of the series reactor at the arc furnace increased the inductive impedance of the furnace and decreased the power factor of the furnace plant. The average power factor of the Davsteel furnace was 0,823 without a series reactor and decreased to 0,774 when the series reactor was installed.

The decrease in power factor ensures that the furnace operates nearer to the sinusoidal supply voltage peak when the current crosses zero. The arc stability is consequently increased, allowing for an increase in arc voltage and arc length. The changes made to the furnace can be seen on the S-locus in figure 5:

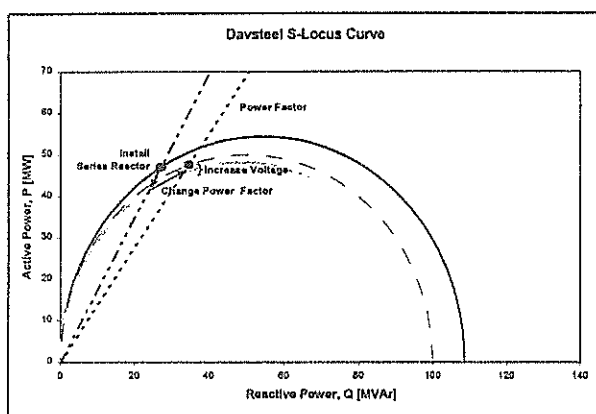


Figure 5: Adjustments made to the S-locus of the Davsteel 60 MVA arc furnace

The series reactor increased the impedance of the furnace's electrical system and decreased the S-locus of the furnace. The active power input into the furnace will be reduced if the furnace maintained operation with the same power factor. The power factor must be reduced to allow more active power to flow into the furnace. To obtain the same active power flow into the furnace with the series reactor, the voltage must also be increased to re-adjust the S-locus of the furnace. This voltage increase can be obtained by increasing the supply voltage or by changing the transformer tapping protocol of the furnace.

To evaluate the impact of the new series reactor on the flicker emission from the arc furnace, flicker measurements were conducted during two measuring campaigns. The first measurement campaign reflects the furnace operation without series reactor and without power factor correction (9,7 MVar) and the second campaign the furnace operation with the new series reactor and with power factor correction. The short circuit power of the Davsteel 60 MVA arc furnace was 125,3 MVA before the series reactor was installed and 100,9 MVA after the series reactor was installed (assuming same voltage conditions).

The higher arc stability obtained with the series reactor, in combination with a slight supply voltage increase due to the power factor correction, lead to an increase of active power input resulting in a 3,6 % decrease of melt-down time and a 4,5 % decrease in flicker Pst 95 %.

4.2 Operation of the open arc furnace

Davsteel operates an open arc furnace where the electrode tips are not covered by charge material during the production process. 55 Ton and 34 ton clamshell baskets are used to load the charge material into the furnace. The charge-baskets are charged with the material to be melted before emptying the basket in the furnace.

Shredded scrap and light scrap is loaded into the bottom, ± 30 %, of the basket and covered with sponge iron, limestone dolomite and anthracite. Heavy scrap and cast iron is then loaded into the basket, that is 50 % full and large enough (55 ton), before it is filled to capacity with general steel scrap and more shredded scrap.

A typical production cycle consisted of two baskets loaded into the furnace to supply the maximum capacity of 75 ton charge material that may be melted in the furnace. Once the charge is in the furnace, it is visibly mixed and fills the whole furnace. If the charge material pile is higher than the furnace roof, the charge crane normally rams it down with the charge basket.

The Davsteel arc furnace employ oxygen-fuel burners to increase the energy input into the furnace. During the first basket, the burner blows at a rate of 600 m³/h for 11 minutes before reducing to 255 m³/h until the second basket is loaded. After the second basket is loaded, the burner blows at a rate of 600 m³/h, for 4 minutes, before reducing the flow rate to 255 m³/h for the rest of the production cycle.

The production cycle makes full use of the available taps on the furnace transformer to limit the power flow into the furnace during the start of each melt. Since the period when the melt is started is the most critical flicker emission period, the implementation of this regulation effectively reduces the magnitude of current fluctuations and flicker from the furnace.

The AC arc-furnace employs three electrodes that are connected to the supply through a furnace transformer. Power flow into the furnace is controlled by adjusting the arc length and arc voltage, while the impedance in series with the arc, is adjusted by using different tapping points on the furnace transformer [7]. The arcs that are established by the electrodes exert electromagnetic forces to repel each other and are very unstable [7].

Furnace current fluctuations are caused by all the factors influencing the impedance of the arc. This impedance depends on the length of the arc and the gas in which the arc is formed [2],[7], [8]. Short-circuits, between two electrodes or between one electrode and the bath, will result in an increase in supply current and flicker emission [4].

The short circuits that may occur in the furnace depend on the volumetric size of the charge material, the control philosophy employed at the arc furnace and the mechanical stability of the charge material.

5 Analyses of results

5.1 Flicker emission coefficient

A histogram provides an accurate tool to differentiate between the flicker emission from the arc furnace and that of the ladle furnace. The histogram of flicker emission, calculated for every heat recorded during the two measurement campaigns, is shown in figure 6. This histogram does not represent the furnace when the flicker emission level is low, but shows the influence of the ladle furnace on the flicker emission of the whole system.

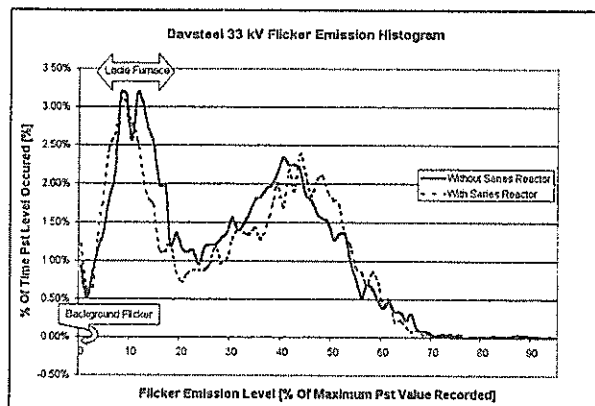


Figure 6: Histogram of Davsteel 60 MVA furnace flicker emission coefficients

The installation of the series reactor only slightly decreased the average flicker emission Pst 99 % due to the increase in the magnitude of current fluctuations that cause flicker. The daily flicker emission coefficients, Kst's shown in table 1, were calculated with equation (13) and equation (14) that are based on the measured Pst 99 % flicker level.

The Kst value of 76,3 measured during the whole period for the furnace with the series reactor installed, is only marginally higher than the proposed Kst design level of 75 [5]. The introduction of the series reactor resulted in an increase of 18,7 % in the Kst value for the whole period. The flicker performance of the furnace is considered to be more predictable with the series reactor.

The adjusted flicker emission coefficients, Kst_{Adjust}'s, are higher than the normal flicker emission coefficients, Kst's. The maximum Kst_{Adjust} of the furnace without series reactor is 15 % higher than the maximum Kst level of the furnace without series reactor. The maximum Kst_{Adjust} level increased by 6,1 %, from 83,7 to 88,8, compared to the 17,0 % increase when Kst values are used. A Kst_{Adjust} level of 83,7 is proposed to represent the furnace without series reactor and a Kst_{Adjust} level of 88,8 to represent the furnace with series reactor.

Table 1: Daily flicker emission coefficients, Kst's and Adjusted daily flicker emission coefficients, Kst_{Adjust}

	Without Ser. R		With Ser. R	
	Kst *	Kst _{Adjust} #	Kst *	Kst _{Adjust} #
Highest	72,5	83,7	84,8	88,8
2nd	68,8	78,5	82,9	86,6
3rd	67,6	78,0	81,0	84,9
4th	67,6	78,0	77,3	81,0
5th	65,8	75,9	76,6	80,3
6th	64,6	74,6	76,6	80,3
7th	63,0	72,6	75,9	79,6
8th	60,0	69,2	75,9	79,6
7th	59,8	69,0	75,9	79,6
10 th	58,4	67,4	73,3	76,8
11 th	57,8	66,7	73,2	76,7
12th	57,1	65,8	70,9	74,2
Whole Period	64,3	74,2	76,3	79,9

* Based on a fault level of 532 MVA at the Davsteel 33 kV Substation, a furnace short circuit impedance of 3,268 mΩ / phase for a 50 MVA furnace transformer at 639,9 V and 2,1 Ω / phase series reactance at 33 kV.

Based on a maximum apparent power loading of 108,6 MVA for the furnace without series reactor and 96,3 MVA for the furnace with series reactor.

The maximum measured apparent power S_{max} for the furnace without reactor is smaller than the calculated short circuit power of the furnace, S_{ccf}, without reactor (125,3 MVA). The maximum measured apparent power, S_{max}, for the furnace with reactor is closer to the calculated short circuit power of the furnace, S_{ccf}, with reactor (100,3 MVA). The S_{max} values seem to be a reasonable approximation of S_{ccf}.

5.2 Flicker emission level

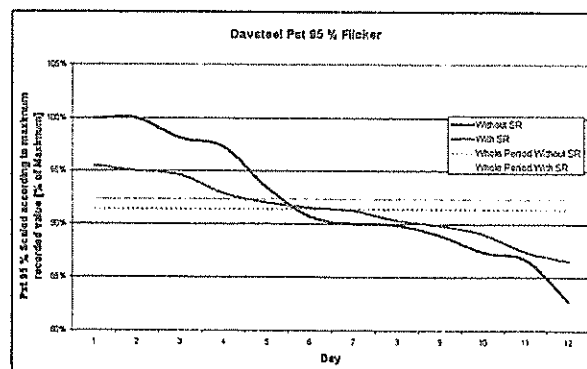


Figure 7: Pst 95 % Flicker level for the furnace with and without the series reactor

Disregarding the flicker mitigation properties of the series reactor, the flicker emission performance of the furnace, expressed in terms of Kst, is worse with the

series reactor installed. Reductions in the Pst 95 % flicker emission levels are, however, evident due to the current stabilizing effect of the series reactor, as is shown in the day distribution curve for Pst 95 % values of the furnace, with and without series reactor, in figure 7.

A clear reduction in the Pst 95 % fluctuation band is evident from figure 7. The reduction in the Pst 95 % fluctuation band resulted in a higher Pst 95 % value, measured over the whole period, and a corresponding reduction in the maximum daily Pst 95 % flicker emission level. The Pst 95 % flicker level can be considered to be more stable after the series reactor was installed.

The 4,5 % reduction in the maximum daily Pst 95 % flicker emission level will only be marginally helpful to industrial furnace operators in respect of the South African NRS 048 electrical standard. This standard specifies that the maximum Pst 95 % from all sources may not exceed a value of 1 during any day of a seven day measurement period at the PCC (88 kV).

5.3 Charge material influence on flicker

The furnace charge consists of general steel scrap, bales, shredded scrap, cast iron, sponge iron, burnt lime, limestone, anthracite and dolomite to be melted during the production process. General steel scrap refers to any piece of steel waste excluding shredded pieces of steel and bales. This scrap may have any size smaller than 0,6 m. Internal Davsteel scrap as well as scrap bought from external suppliers may be classified as steel scrap. A bale consists of steel scrap that was compressed into a rectangular block. This block will have typical dimensions of 0,65 m x 0,4 m x 0,4 m. Bales are not always available and are preferably loaded whenever they are obtained. Cast Iron is obtained from a reduction furnace that refines iron ore to iron. The cast iron is very rich in carbon and must be refined further to produce steel.

Shredded scrap consist of steel scrap that was cut into pieces smaller than 0,075 m. This scrap is a lot lighter than general steel scrap and bales. Sponge iron consist of directly reduced iron pellets with sizes smaller than 0,02 m and a total iron (Fe) content between 87,6 % and 90,8 %. The metalisation of sponge iron is: between 88,5 % and 92,1 %.

To analyze the influence of charge material size on the flicker emission from the furnace, general steel scrap, bales and cast iron was grouped as large material. A percentage point was then used to express the relation between the weight of the large material and the total weight of charge material in the basket. The specific weight influence of the material was accordingly balanced with the influence of other materials loaded into the furnace.

The spread of Kst values obtained during each melt, in relation with the percentage of basket weight consisting of large material, is shown in figure 8.

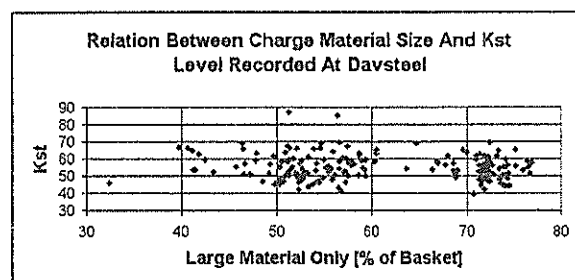


Figure 8: Charge material size relation with the measured flicker emission coefficient

A ten point, or ten melt, sliding average of Kst values are shown in figure 9. This figure employs maximum and minimum indicators to present the range of possible Kst values and are based on the distribution of percentage large material weight.

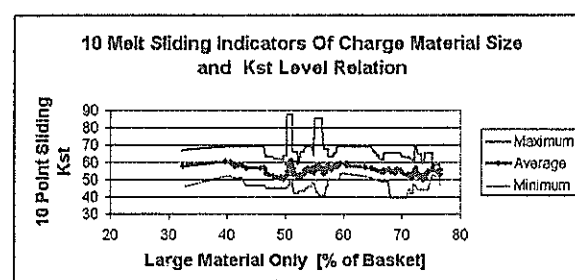


Figure 9: Charge material size relation with statistical indicators for the flicker emission coefficient

The sliding average of Kst values decrease slightly as the amount of large material loaded into the bath is increased. The range of possible Kst values is more stable for higher percentage loading of large material into the bath.

The ten melt sliding window statistical indicators of the average Pst values measured for each melt is shown in figure 10. Again the ten melt sliding average, maximum and minimum indices are used to provide a range indication of possible average Pst values.

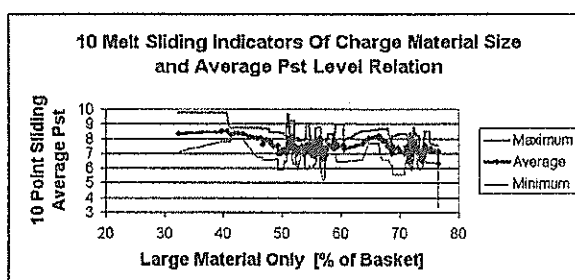


Figure 10: Charge material size relation with the measured flicker emission coefficient

A decrease in the amount of average flicker generated by the furnace can be detected when the percentage large material loaded into the basket is increased. This confirms that the charge material size does influence the average flicker emission characteristics of the arc furnace.

5.4 The influence of the charge material loading technique on flicker

The furnace charges material into the furnace by swinging the roof open to allow a crane to empty a charge basket into the furnace. A 55 ton charge-basket is loaded into the furnace to start a melt and a 34 ton charge-basket is added to the melt within the first 15 minutes of a melt. The charge sizes are controlled very well and large material is never loaded into the second basket.

The flicker emission coefficient was determined to correspond to the flicker level recorded during each basket. The first basket is represented by the flicker measurement data that was recorded from the onset of the melt until the second basket was loaded. The second basket is represented by the flicker measurement data that was recorded as soon as the second basket was loaded until the melt was finished. The data for both baskets were filtered to ignore all flicker levels below 0,75.

The flicker emission results shown in table 2 were obtained with the corresponding flicker emission coefficient.

Table 2: Comparison of the flicker emission prediction coefficient results obtained for the two charge baskets used at Davsteel

Basket	Flicker @ Munic Sub		Kst	Kst Adjusted
	Pst Max [#]	Pst 95 % [#]		
1: 55 Ton	2,88	2,07	68,4	79,2
2: 34 Ton	2,56	1,92	65,3	75,7
Difference			4,5 %	

Based on flicker emission measurements at the 33 kV / 11 kV furnace transformer.

The smaller basket was loaded after the first basket was already melted. The Pst 95 % flicker level for the second basket was 7,1 % lower than the flicker level of the first basket and the flicker emission coefficient of the second basket was 4,4 % lower than that of the first basket [10].

The furnace is generally loaded to the full when both the first and the second basket are loaded, but exceptions occur when the second basket is loaded. Large material is only loaded into the first basket and it is assumed that the material loaded with the first basket is more densely packed than the material loaded with the second basket. A molten bath is left in the

furnace, when the furnace is tapped, to heat the charge material of the first basket and to join the charge when the molten metal solidify. The furnace contains a larger molten bath when the second basket is loaded.

The highest flicker peaks occurred when the first basket was loaded and the factors described above may cause a reduction in the flicker levels between the first and second basket. The flicker peaks are caused by disturbances within the furnace that perturb the electrical arc.

6 CONCLUSIONS

The installation of the reactor should theoretically lead to a 19,5% reduction of the furnace short circuit power if the same voltage is used at the plant medium voltage bus-bar and the same transformer ratio is used at the furnace transformer. A consequent reduction of flicker corresponding to approximately the same amount is expected. The operating conditions of the furnace changed, due to (i) a shift to a lower load-factor, (ii) an increase of voltage due to power factor correction and / or (iii) changes of the furnace transformer ratio, to allow the series reactor to attain an increase of the average power input to the furnace and a reduction of melting time without increasing flicker emission

The flicker emission coefficient Kst, calculated on the basis of the theoretically reduced short circuit power Scsf and the measured nearly unchanged flicker, consequently increased by 18,7 % to 76,3. This value corresponds better with published Kst - values and is more predictable. The series reactor smoothed the Pst 95 % level experienced at the furnace previously, resulting in a slight decrease in the Pst 95 % level of around 4,5 %.

A decrease in the flicker emission level was detected when the percentage loading of large material was increased. This affirms that the charge material size does influence the flicker emission from the furnace. It is interesting to note that a higher share of small and light material apparently leads to higher flicker because small parts are more easily tossed away by the arc, thus, causing larger and more arc length variations.

7 ACKNOWLEDGEMENTS

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availability of their arc furnace for this study is deeply appreciated.

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NEURO-FUZZY NETWORKS: A BOON TO CONDITION MONITORING

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Abstract: In modern industry economic pressures dictate that unnecessary outages due to maintenance should be kept to a minimum. This increases the need for online condition monitoring systems to assess the health of electrical machines. A condition monitoring system rests on two pillars, viz. non-invasive measurement techniques and the interpretation of data. Artificial intelligence based classifiers can play an important role in automating the diagnostic process. In designing a classifier, it would be invaluable if the experience of human operators in determining the condition of a machine could be utilised during the training process. Fuzzy logic based-classifiers are quite adept at precisely this task. However, fuzzy logic systems are hamstrung by the fact that they are only optimally calibrated with extreme difficulty. Fortunately neural networks exhibit the ability to "learn" quantitative relationships. By combining neural networks and fuzzy logic it is possible to retain the best of both worlds. Consequently neuro-fuzzy classification networks should be given serious consideration in the design of a condition monitoring system.

Keywords: neuro-fuzzy networks, classification, condition monitoring

1. CONDITION MONITORING

Previously, the mainstay of any factory's maintenance program was preventive (or time-based) maintenance. Basically, this entails taking a machine out of active duty at regular intervals and performing a complete manual inspection [1]. The main point of criticism against preventive maintenance is that inspections are carried out regardless of the machine's actual condition. This means that more often than not, a machine is still in mint condition when it is taken out of production for an inspection. Consequently preventive maintenance leads to unnecessary loss in production, increased labour costs and sometimes even to damage to the machine [1], [2].

Advances in the fields of computer architecture and pattern recognition have led to the birth of condition-based maintenance (alias predictive maintenance). Condition monitoring is the process by which the condition of a machine is estimated on a continual (and non-invasive) basis *whilst it is in active duty*. Because machines are only removed from operation when more intensive tests or visual inspection are needed to confirm a faulty condition diagnosed by the condition monitoring system, outages and labour costs are reduced dramatically [1]. A prime example of condition monitoring is the identification of partial discharge signals present in an active machine, with the aim of assessing the condition of the machine's insulation system.

One important element in a condition monitoring system is an automatic classifier whose task it is to correctly identify the condition of a machine based on inputs from sensors. An interesting aspect of condition monitoring is that the phenomena which are being monitored (e.g. partial discharges) have themselves an impact on the condition of the machine [3].

Consequently, the boundaries between different classes of conditions aren't clear-cut.

Besides being able to identify classes with hazy boundaries, it would also be invaluable if the classifier could incorporate some of the human operator's expertise in assessing the true condition of a machine, since this would greatly lessen the burden on the training algorithm. Luckily fuzzy logic-based classifiers exhibit both of these abilities, which makes them ideally suited for the condition monitoring problem.

2. FUZZY LOGIC-BASED CLASSIFIERS

2.1 Motivation

Conventional classification techniques are based on the assumption of an object's undivided loyalty to a single class. However, a certain amount of overlap between different classes in the feature domain is a fact of life in practical identification problems*. Therefore, fuzzy classification systems are more realistic in that they allow an object to have degrees of membership to different classes [5].

A more insidious problem in traditional pattern recognition systems is the fact that the very training examples, upon which a classifier is trained, may be erroneous. Most commonly, the root cause of this problem is subjectivity or inconsistency on the part of the human expert responsible for compiling the training set [7]. In this case fuzzy logic techniques may be employed to generate more accurate artificial fuzzy class labels [8].

* Reasons for the overlap between classes can be found in inadequate feature extraction procedures, measurement noise, missing measurements but also the fact that signals are sometimes inherently member of more than one class [5], [6].

2.2 Basic concepts

Fuzzy logic is an emulation of certain aspects of the human reasoning process as well as our ability to summarise situations and handle uncertainty in a robust manner [4]. It is based on the notions of fuzzy sets, linguistic variables and fuzzy logic rules.

A clear and unambiguous distinction between those elements belonging to a particular set and those who don't, is the characterising feature of *crisp sets*. In contrast, a *fuzzy set* allows a gradual change between full membership and non-membership by associating a degree of membership with every element in the set [9]. A graphical illustration of this idea can be found in figure 1. This figure consists of a grey-scale image showing the degree of membership to one of the classes in the classical XOR-classification problem (which has been generalised by allowing overlap between the different clusters) upon which representatives of both classes have been superimposed.

Linguistic variables differ from normal variables in that they have a limited amount of possible values, each of which is expressed in terms of words. An example is the linguistic variable: *temperature* which could consist out of the values: *hot*, *cool* and *cold*. The distinct values of a linguistic variable are generally known as its *terms* and are represented by specialised fuzzy sets called fuzzy numbers [9].

Fuzzy rules make extensive use of linguistic variables and are useful for the storage of human experts' knowledge.

Most commonly, fuzzy classification rules take the following general form [9]:

$$\text{IF } (\tilde{A} \text{ is } T_{\tilde{A},1}) \text{ AND } (\tilde{B} \text{ is } T_{\tilde{B},1}) \text{ THEN } (\tilde{C} \text{ is } \tilde{C}_1) \dots (1)$$

In which $\tilde{A}$ and $\tilde{B}$ are linguistic variables representing different features and $T_{\tilde{A},1}$ and $T_{\tilde{B},1}$ are examples of their respective terms. Furthermore, the classification decision is represented by the linguistic variable $\tilde{C}$ consisting of different fuzzy classes (e.g. $\tilde{C}_1$).

Different methods of fuzzy inference can be used to determine (from the available fuzzy classification rules) a particular signal's degree of membership to the different classes. For example, the following procedure can be followed:

- Step 1. Determine the degree similarity of each feature value to its respective terms.
- Step 2. Calculate the degree of applicability (or firing strength) of each classification rule by means of a fuzzy AND operation.
- Step 3. Combine the firing strengths of rules having the same consequences with the help of a fuzzy OR operator in order to determine the degree of membership to the different classes.

2.3 Advantages and disadvantages

It is possible to make use of the specific nature of the outputs of fuzzy classifiers in order to develop a prejudice for or against certain classes. In a condition monitoring application this would enable a fuzzy classifier to avoid false alarms and the resulting financial losses [10].

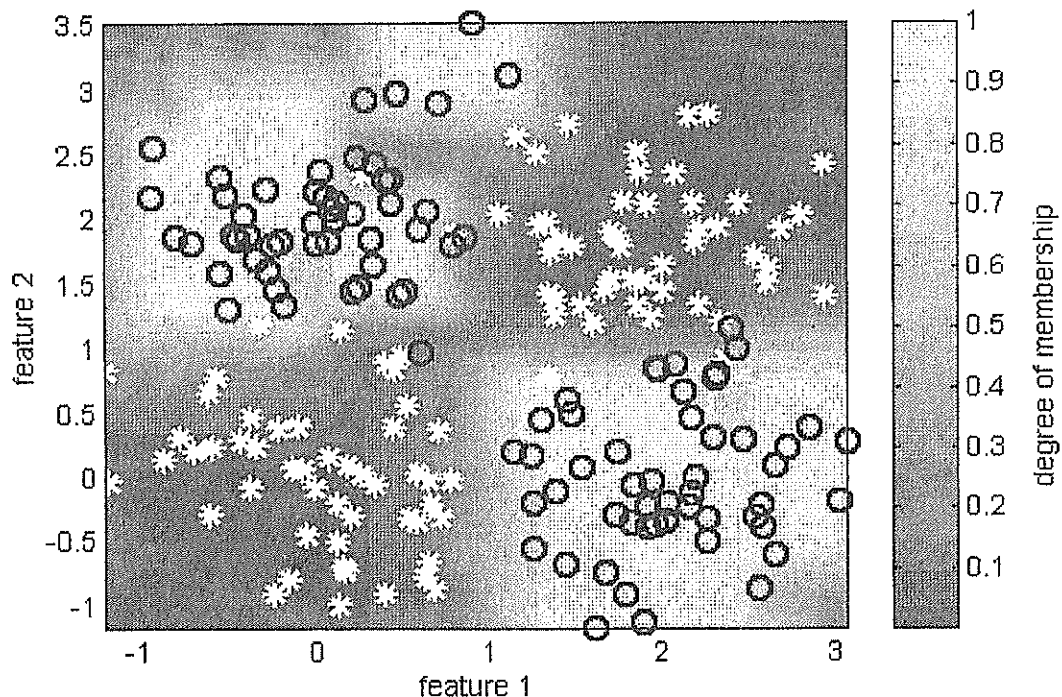


Figure 1 Contour plot of a fuzzy class in the generalised XOR-problem.

Further benefits of the use of fuzzy classifiers include:

- They are capable of handling noisy, imprecise and even missing measurement values in a robust fashion [5], [6].
- The ability of fuzzy systems to interpolate between existing rules (due to the unique nature of fuzzy sets) permits them to operate on far fewer rules than other rule-based classifiers [4].
- With minor modifications a fuzzy classifier can accept linguistic inputs [6]. Furthermore, its outputs are easily interpreted by a human operator [12]. Consequently fuzzy classifiers possess a better interface with humans than many other classification techniques.

However the growth of fuzzy technology has for long been stunted by the difficulty in obtaining fuzzy rules and defining their fuzzy sets for optimal use [4].

3. NEURO-FUZZY NETWORKS

3.1 Neural networks to the rescue

Basically, a neural network is a collection of processing units (called nodes) arranged in a specific pattern and provided with adjustable parameters (also known as weights) on the connections between nodes [11]. The overall input/output relationship for a neural network is determined by its weights and the operations performed in the nodes.

Neural networks are best known for their ability to learn (and to a certain extent generalise) complex, non-linear relationships from numerical examples [14]. Besides being unable to handle certain types of uncertainties, their main drawback lies in the similarity between their method of operation and that of a black box [12].

3.2 A marriage of complementary techniques

By comparing neural networks to fuzzy logic systems,

it can be concluded that they are complementary techniques with the disadvantages of each being compensated for by the advantages of the other [4]. Consequently a combination of these two techniques could possibly deliver dividends. In one such a hybrid classifier the neural network serves as an implementation and training framework in which the consecutive layers of the network correspond to different stages in the fuzzy inference process. These hybrid systems go by the name of *heterogeneous neuro-fuzzy networks*, an example of which is shown in figure 2.

The most important contribution of neuro-fuzzy networks is their ability to extract fuzzy rules from numerical training examples [13]. Accordingly, neuro-fuzzy networks aren't nearly as dependent on the guidance of a human expert as is the case with conventional fuzzy systems. Furthermore, their rule generation capability makes neuro-fuzzy networks a potentially important knowledge acquisition tool [12].

Due to their rule-based functioning, neuro-fuzzy networks' internal structure are generally much more understandable than that of conventional neural networks [12]. Hence it is comparatively easy to incorporate a human experts' experience in the structure of a neuro-fuzzy network.

However, neuro-fuzzy networks shouldn't be regarded as the panacea for all classification-related problems, because they too have a few shortcomings. Like most classifiers they are dependent on a good selection of features describing the signal to be identified, in order to produce good classification results. Furthermore, their training procedures (responsible for extracting rules and refining their parameters) can get quite complex and computationally expensive.

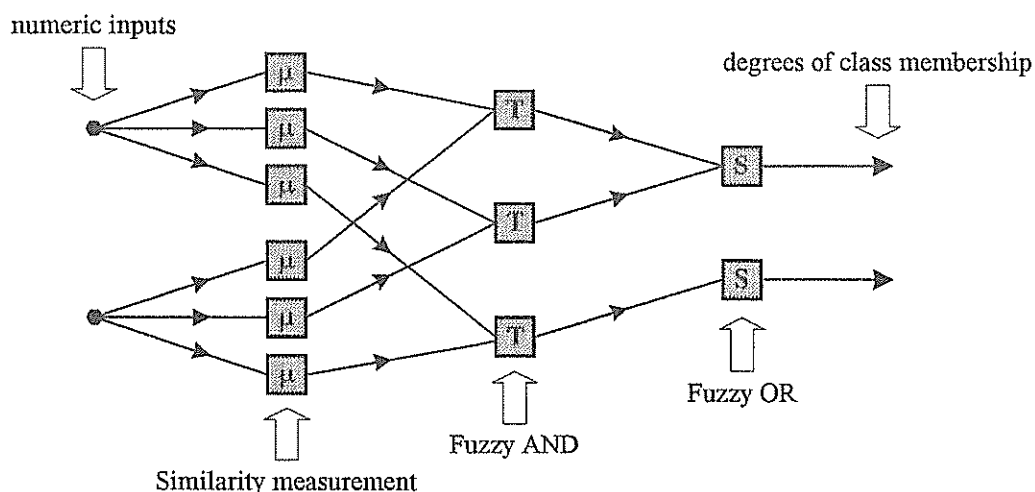


Figure 2 Schematic diagram of a heterogeneous neuro-fuzzy network.

3.3 Incorporation of human experience

It has been previously mentioned that a neuro-fuzzy network can be enriched by the inclusion of a human expert's knowledge. The first step in this process is that the expert's experience be cast in a manageable form (e.g. a flowchart). An example of such a flowchart (for the specific case of partial discharge monitoring) can be found in [15].

Secondly, the different input- and output variables should be identified and redefined in terms of linguistic variables with associated linguistic values (terms). At this stage only a tentative specification of each term's fuzzy set is necessary. Armed with these definitions, fuzzy rules have to be extracted from the flowchart (or general rules) obtained in the first step. The resulting fuzzy rule-base serves as a starting point from which an automatic rule-generation algorithm can deliver optimised fuzzy rules, with the assistance of numerical training examples.

For the sake of illustration, the following "rule" was extracted from the flowchart in [15]:

If $150 < Q_m < 600$ and $+Q_m/-Q_m < 0.5$ and a negative temperature effect[†] is observed, then significant copper separation from the groundwall has occurred.

In the above rule Q_m represents the peak partial discharge (PD) magnitude at a specific pulse repetition rate, whilst $+Q_m/-Q_m$ is the ratio of the peak positive PD magnitude to the peak negative PD magnitude.

By defining the linguistic variables: *peak PD magnitude*, *PD magnitude ratio*, *temperature gradient*, *load gradient* and *PD gradient* and providing each with a few linguistic values (e.g. *low* and *high*), the following fuzzy rule has been obtained [16]:

IF (peak PD magnitude is medium) AND (PD magnitude ratio is low) AND (load gradient is small) AND (temperature gradient is negative) AND (PD gradient is high) THEN significant copper separation from the groundwall has occurred.

4. CONCLUSION

Neuro-fuzzy networks exhibit abilities which should make them powerful techniques to tackle real-world classification problems. They should be particularly effective in instances where noise and other uncertainties result in significant overlap between classes in the feature domain. The ease with which neuro-fuzzy networks can both absorb human knowledge and generate linguistic rules from numerical examples, should almost make them

compulsory techniques when a task previously performed by a human operator has to be automated.

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[†] A negative temperature effect is defined by at least a twofold increase in the peak partial discharge magnitude as the temperature is lowered from a high to a low level (with the provision that the electrical load remains unchanged).

STRATEGIES FOR DAMPING POWER OSCILLATIONS USING CONTROLLABLE SERIES COMPENSATORS

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Abstract: This paper provides a review of the control strategies proposed for damping power system dynamic oscillations using controllable series compensators. The applicability and suitability of these control strategies for various types of power oscillation damping is also analysed.

Keywords: Power Oscillation Damping, Flexible AC Transmission Systems, Controllable Series Compensation.

1. INTRODUCTION

A steady growth in load demand coupled with the cost of land required to accommodate new power lines, limited availability of right-of-ways, regulatory policies, and the cost of building new generation plant to meet this increased power demand will inevitably require better utilisation of existing generation and transmission facilities. This in turn implies the need to operate the existing power generation and transmission system much closer to its stability limits without compromising the security of the system.

The response to these pressures, internationally, has been the emergence of the FACTS (Flexible AC Transmission Systems) concept: FACTS aims to bring the factors which influence the transfer of power in a transmission system under rapid control by means of modern, power-electronic-based compensating devices in order to allow the system safely to be pushed closer to its limits. Fig. 1 shows a simple diagram of a two-area power system connected by a transmission corridor. The power P_{tr} transferred down the line reaches a maximum $P_{trmax} = |V_S||V_R| / X_L$ when the transmission angle δ reaches 90 degrees, where $|V_S|$ and $|V_R|$ are the transmission system voltage magnitudes and X_L is the series (inductive) reactance of the transmission line. This maximum power transfer P_{trmax} is the *steady state* limit of the system. However, the dynamic and transient stability characteristics of the synchronous generators in the system mean that the transmission line must in practice be operated at a power level much lower than this steady state limit in order to allow sufficient margin for the generators to return to synchronism following disturbances. FACTS devices aim to enhance the control of the electromechanical power oscillations between the generators and the transmission system so as to reduce the need for such large transient and/or dynamic stability margins.

From Fig. 1 it is apparent that three possible levers exist for control of power flow: voltage magnitude, voltage phase and line reactance; however, this paper is concerned with one particular lever, namely the series reactance of the transmission line. Traditionally, fixed

capacitor banks have been used in series with transmission lines to reduce their net inductive reactance and thereby to increase their maximum power transfer capability: with a capacitive compensating reactance X_C introduced in series with the line of Fig. 1 the steady state power transfer limit is increased to $P_{trmax} = |V_S||V_R| / (X_L - X_C)$. Modern, FACTS series compensators now allow such series compensating reactance to be controlled dynamically. The purpose of this paper is to consider, in particular, the issue of how best to control such a dynamically variable series compensating reactance in order to damp out the electromechanical transients in a power transmission system.

2. HISTORICAL PERSPECTIVE

In 1966, E. W. Kimbark [1] proposed the use of a switched series capacitor to change the power flow following a disturbance. He used equal area criterion to predict that the transient stability of an electrical power system can be improved by inserting and bypassing a series capacitor. The advantage of switched versus fixed series compensation is that it increases both the steady-state and transient stability limit allowing the system to be operated closer to theoretical stability limits, and it is more economical. It must be noted that the analyses in [1] referred to one switching in of a series capacitor after a fault. In 1969, O. J. Smith [2], proposed that the electromechanical power oscillations between a generator and a transmission system can be damped by inserting and removing a series capacitor in a transmission line. While reference [1] proposed a single switching in and out of a series capacitor, Smith appears to be the first to propose a coordinated insertion and removal during the transient power swings. Reference [2] states that even a small amount of switched series compensation can provide a dramatic increase in the power system stability. In a subsequent paper by Smith [3], the ideas in [2] were practically implemented in the laboratory and it was shown that a generator which would otherwise have pole-slipped following a particular system fault was successfully stabilised using switched series capacitance.

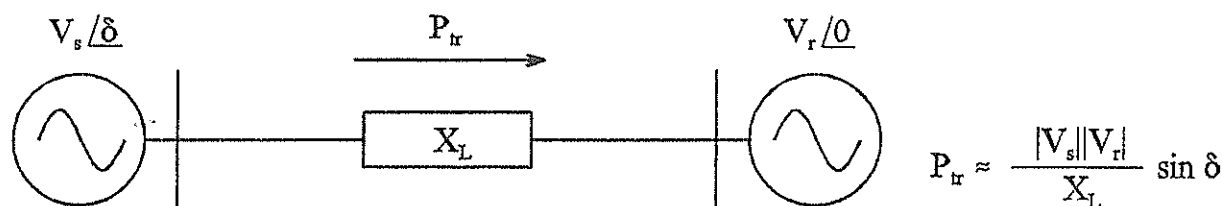


Fig. 1 - Simple two-area power system illustrating the factors which influence power flow.

At the time the ideas in [1-3] were proposed, dynamically variable series compensation could only be achieved using mechanical circuit breakers; however such devices are too slow and unreliable for the repeated high-speed use demanded by a dynamic control application and as a result these ideas remained on hold for nearly two decades. Recently however, progress in the field of power electronics has led to the development of high-power electronic switches with which to implement dynamically controllable (FACTS) series compensating devices. A number of different FACTS series compensators have subsequently emerged, based either on the thyristor (eg. the Thyristor Controlled Series Capacitor or TCSC) or on GTO inverters (eg. the Static Synchronous Series Compensator or SSSC), and any of these FACTS series compensators can now, in principle, be used to provide a dynamically varying series compensating reactance in a transmission system for the purpose of enhancing system stability. The following section reviews the particular methods which have been proposed for controlling such dynamically varying compensating reactances for the specific purpose of damping of electromechanical power oscillations.

3. CONTROL STRATEGIES FOR DAMPING ELECTROMECHANICAL POWER OSCILLATIONS.

As mentioned previously, Smith [2] appears to be the first to propose multiple and timed insertion and removal of a switchable series capacitor to damp the post-disturbance electromechanical oscillations of a synchronous generator. Smith considered a two-generator system, and proposed to insert the series capacitor in an interconnecting transmission line at the beginning of the transient and remove it when the derivative of the power flow in the line becomes zero; a further insertion was proposed if the kinetic energy due to the generator's speed deviation matched the amount of kinetic energy that could be removed during the rest of the swing. A second control law was also proposed in [2] whereby immediate insertion of the capacitor occurs after the disturbance and its first removal occurs when $d\delta/dt < 0$; subsequent re-insertion and removal is based on a decision function.

The control laws proposed in [2] aim to return the generators to steady state in a dead-beat (non-oscillatory) manner, but they require the use of complicated decision functions which in turn require a

knowledge of system angular frequency, shaft angular velocity, moments of inertias of the two machines, target steady-state power flow, reactance of the series capacitor and the voltage behind quadrature axis reactance in each machine. The measurements required for the controller are machine angular velocity w , power P_{tr} , a knowledge of which line has been lost as well as pre-transient information. Due to its complexity, the control proposed in [2] is likely to be impractical; indeed the author himself adopts a far simpler and more practical control approach in his subsequent paper [3].

Smith and Webster [3] describe a practical, laboratory-scale implementation of the ideas in [2] using a far simpler control law. The control adopted is to insert and bypass a series capacitor based on the deviation $w - w_0$ between generator shaft speed w and synchronous speed w_0 . The proposed bang-bang algorithm is as follows:

insert X_C if $w - w_0 > 0$
bypass X_C if $w - w_0 < 0$(1)

The authors chose the value of X_C using successive trials, but show, by means of an equation, that a maximum value of X_C exists above which negative damping can be introduced. A further, and well-known, problem with this bang-bang approach is that of controller chatter: near steady state, small speed deviations can introduce unwanted operation of the control. The authors avoid this problem by including a time-based, exponentially-decreasing override function: this approach enables the full benefit of the bang-bang control when the system is far away from target (steady-state) and gradually phases out the control as the system nears steady-state.

A number of papers by Swift et al. [4-6] investigate power oscillation damping using controllable series compensators (CSC). In particular, reference [4] investigates the CSC damping control by studying a single machine infinite bus (SMIB) system. The authors integrate a model of the CSC into the well known 'Phillips-Heffron' model of a synchronous generator; a control law is assumed for the CSC of the form

$$\Delta x_{CSC} = \frac{K_C T_{ws}}{1 + s T_{ws}} w_0 \Delta w \quad \dots(2)$$

where the input to the CSC control is the generator speed deviation Δw , the output of the control is the

change Δx_{sc} in the series compensation and K_C is the controller gain. The authors present a detailed analysis to show that the damping provided by the CSC control increases with increasing gain K_C ; the findings of Swift's analysis therefore provide a theoretical justification for the control approach adopted in [3] and shown here in eqn. (1): for a given magnitude of controllable series capacitive reactance, the best improvement in damping is shown to occur for infinite controller gain K_C (ie. bang-bang control). The analysis in [4] also suggests that for robust control performance, the CSC damping control must be designed at conditions of light line loading and weak system connections (the conditions for which the control is least effective). The attraction of the analytical approach of Swift et al. is that it yields insightful design equations for the synchronising and damping torques provided to the system by the CSC's damping control. Indeed, reference [5] extends the ideas in presented in [4] into a multi-machine application and, using the sensitivities of CSC damping torque to the location of the device, is able to determine the optimum location of a CSC controller in a multi-machine environment.

The control strategies discussed thus far have all concentrated on damping the power swings of one or more generators in cases where the behaviour of each individual generator is of direct concern. However, unstable power swings also frequently occur in transmission systems as a result of a large group of generators in one area oscillating against a large group of generators in another, geographically remote, area. Damping such inter-area mode power oscillations using controllable series compensators typically requires a distinct control approach.

Reference [7] examines the control of inter-area mode oscillations using controllable series compensators in a two area system, with each area represented as an aggregate machine; a continuous control law is proposed as follows,

$$\Delta x_{sc} = K_{sc} (\Delta w_1 - \Delta w_2) \dots\dots\dots(3)$$

with the control signal selected as the difference between the angular velocities of the two aggregate machines. In practice it would be necessary to synthesise the signals Δw_1 and Δw_2 from measured variables in each area, although this aspect is not considered in [7]. The analyses in [7] show that in the case of the inter-area mode phenomenon the damping due to the CSC increases with line loading; this is shown to be advantageous since normally under this condition the damping of inter-area modes decreases.

Gronquist [11] considers the problem of power oscillation damping using a TCSC and exclusively locally measurable variables for its control. The control law derived in [11] is based on the volt drop E across the line in which TCSC is placed and its derivative $\dot{E}$.

The voltage E is synthesised by multiplying the measured line current by the known reactance of the line and adding this to the voltage across the TCSC itself as follows:

$$E = I X_{LINE} + E_{TCSC} \dots\dots\dots(4)$$

The input to the controller is then the product of the synthesised voltage E and its derivative $\dot{E}$; based on this input signal a continuous damping-control law is proposed which can be summarised as

- $E \cdot \dot{E} > 0$: increase TCSC reactance
- $E \cdot \dot{E} < 0$: decrease TCSC reactance
- $E \cdot \dot{E} = 0$: keep TCSC reactance at nominal.

Similarly, de Mello [13] proposes that in the case of controllable series compensation, continuous control using locally measured variables is the appropriate control approach for inter-area mode damping control; in particular de Mello proposes using the rate of change of the power flowing through the controllable series compensator itself as the most logical input control signal for power oscillation damping of inter-area modes.

4. DISCUSSION

The previous section has briefly reviewed the control approaches which are proposed in the literature for power oscillation damping using controlled series compensators. These approaches may be categorised into two broad groups.

Bang-Bang Control

Several authors have shown bang-bang control of the controllable series compensating element to be the most efficient method of increasing the damping of the electromechanical oscillations of the synchronous generators in a power system. Despite the theoretical advantages of this approach however, it has a considerable disadvantage in the case of small disturbances, where the response of the control to such disturbances may itself excite undesirable system behaviour. Although the analytical techniques of Swift et al. [4-6] are able to identify the synchronising and damping torques obtained using continuous CSC control, no definitive method has been proposed for designing the magnitude of controllable series compensating reactance which should be employed in a bang-bang regimen of control. The bang-bang method of controlling series compensating reactance is thus most suitable for use only in the short time period following a severe system disturbance in order to improve the first swing stability of the generators feeding the transmission system.

Continuous Control

Although continuous control of the series compensating reactance has less impact on power swings it is more suited to the requirements of damping inter-area mode

oscillations, since these oscillations are in effect a small-signal, dynamic stability problem. The additional challenge in using controllable series compensation to damp inter-area modes is the selection of a suitable input signal to the controller; however, as the discussions of the previous section have shown, such signals can be obtained in practice. The advantage of the continuous control approach to damping inter-area mode oscillations is that powerful, small-signal linearised analysis methods (eigenvalues, eigenvectors participation factors etc) may be used to great assistance in designing and locating the controlled compensator.

4. DISCUSSION

This paper has presented an overview of the strategies suitable for damping power system oscillations using controllable FACTS series compensators. The discussions have shown that the choice of control regimen should be guided by the nature of the power oscillation phenomenon to be damped: discontinuous (bang-bang) control of series reactance is suited to maintaining system stability in the first swing following a major disturbance, whilst the various forms of continuous control of series reactance are suited to damping inter-area mode oscillations. The authors are currently researching the use of inverter-based FACTS series compensators for inter-area mode damping, and future publications by the authors will demonstrate the application of some of the techniques described in this paper to that particular problem.

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A HARMONIC RESONANCE CASE STUDY USING A PC BASED SOFTWARE PACKAGE

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ABSTRACT

This paper describes a power system harmonic resonance case study using a PC (personal computer) based software package as the main analytical tool. The network used for the study is a self-designed power factor (pf) corrected 11kV plant. Its parameters have been selected in such a way to cause the harmonic resonance point to coincide with the seventh harmonic generated by an ideal six-pulse convertor and in this way make the study meaningful and lead the investigation into a basic filter design, thus shifting the resonance point out of the range of the lowest harmonic generated by the convertor. The paper describes step-by-step the methodology needed to be followed when doing such a case study using the given software package. The case study applies the methodology making it also useful as a power system educational tool for understanding harmonic resonance in pf corrected plants.

1. INTRODUCTION

Traditionally pf correcting banks of capacitors are added to plants to improve pf's and reduce billing. This may improve the pf problem but in the process it may create other serious problems if there are harmonics present. At present many pf corrected plants are subjected to the injection of $6k\pm 1$ harmonic currents produced by a six-pulse convertor [1].

Any capacitance (C) and inductance (L) form a circuit, which is tuned to what is called its "resonant frequency". This is the frequency at which its capacitive reactance equals its inductive reactance. If this circuit is excited by a signal that is at or near its resonant frequency, it oscillates causing circulating currents, which are much greater than the exciting signal. If sufficient harmonic energy is present and its frequency matches that of the resonant circuit formed by the pf correcting capacitor and the inductances present it will excite the circuit into oscillation. A very large harmonic current will flow producing high voltages that could damage circuit elements.

An obvious solution is to shift the resonant frequency so that it does not coincide with an exciting harmonic. The adding of filters is one way of lowering frequency resonances. Filters are usually effective if added near the loads that are the source of harmonics [2].

There are basically two types of resonance. Parallel resonance presents a high impedance to the flow of current whose frequency corresponds to the resonant

frequency, whereas a series resonant circuit is a low impedance [3]. If the parallel resonant frequency occurs at or near one of the harmonics produced by the load equipment, severe amplification of harmonic voltages and currents will result [4].

A goal of this case study is to demonstrate how a software package can be used to analyse harmonic resonance in a pf corrected plant and how to overcome the resonance problem caused by a six-pulse convertor.

A further goal is to put in place a methodology which will not only demonstrate how to apply a software package for such case studies but which can also be useful as an educational tool for training engineering personnel in power system harmonic resonance analysis techniques.

The software package used for this analysis was the industrial grade ERACS package [5].

2. NETWORK

Figure 1 shows the one line diagram of the resultant 11kV network, including the tuned filter bank.

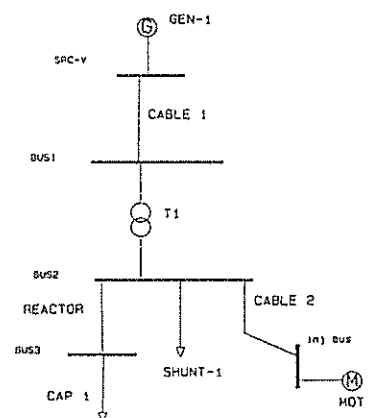


Figure 1: Network

3. NETWORK DESCRIPTION

The network consists of a 11kV generator (GEN-1), representing an infinite bus. The generator is connected to the SRC-V bus, which is connected to bus 1 via Cable 1 and represents an underground cable feed to the plant substation. Cable 1 is modelled as a paper-insulated cable with

aluminium conductors [6]. A two winding transformer (T1), 11kV/3.3kV is connected between buses 1 and 2. A motor (MOT1) is connected to the injection bus (represents convertor) which in turn is connected via a low resistance cable (Cable 2) to bus 2. Bus 2 also has connected to it a shunt load (Shunt1) and a filter. The filter comprises a low resistance reactor and a capacitor (Cap 1) in series.

4. METHODOLOGY

The method used to analyse harmonic resonance in a pf corrected plant and how to overcome the resonance problem caused by a six-pulse convertor is as follows:

1. Model the network in accordance with the parameter requirements of the ERACS package, generate a one-line diagram for the network (figure 1) and perform a loadflow calculation (LF). The package shows modelling errors or if any device is overloaded.
2. Once a successful LF has been completed, carry out an ERACS harmonic impedance scan (H.I.S).
3. From LF results determine by hand calculation the pf from the power, MW (P) and var, Mvar (Q) values for bus 2. Then determine therefrom the Q and C (μF) value of the capacitor bank required for the corrected pf. Model the capacitor bank as a shunt load with a negative var value.
4. Re-run LF to determine if the desired pf has been obtained.
5. Perform a H.I.S to ascertain how the capacitor bank alters the resonant point at bus 2. After pf correction the resonant point usually lies between the 5th and 7th harmonic for pfs > 0.95 [7].
6. Calculate by hand the reactance value for the reactor required in series with Cap 1. Due to pre-selected system parameters the resonant frequency coincides with the 7th harmonic produced by the six-pulse convertor. This resonant point needs to be shifted below the lowest harmonic frequency of the six-pulse convertor. This is achieved by adding a reactor in series with the capacitor bank creating a tuned filter circuit. Such a circuit brings about two resonant points, a parallel (f_p) and a series (f_s) resonant frequency point. The method used to determine the reactor value is as follows [7]:
 - (a) Determine the f_s to which the filter circuit will be tuned. It is recommended to shift f_s to a value 3% to 10% less than the lowest-order harmonic produced by the load equipment (six-pulse convertor)[4]. This is calculated by hand as follows (e.g. using 10%):

$$f_s = f_{5th \text{ harm}} - (0.1 \times f_{5th \text{ harm}}) \dots \dots \dots (1)$$

$$\text{i.e. } f_s = 250\text{Hz} - (0.1 \times 250\text{Hz}) = 225\text{Hz} \\ = 4.5^{th} \text{ harmonic}$$

- (b) Calculate by hand an inductance value for the reactor of the tuned circuit at f_s using:

$$L_F = (1/2.\pi.f_s)^2/C \dots \dots \dots (2)$$

- (c) Calculate the inductive reactance value for L_F at the fundamental frequency (f_0) using:

$$X_L = 2.\pi.f_0.L_F \dots \dots \dots (3)$$

7. Model the tuned filter circuit. In ERACS this is best done by modelling the reactor (X_L) as a line with a small resistance in series with the capacitor bank.
8. Perform a LF and check that the pf value and the voltage across the capacitor bank is not excessive. When a reactor is added in series with a capacitor the voltage across the capacitor rises above nominal voltage [8]. It is standard practice to rate pf correction capacitors at 10% above nominal voltage [9]. The pf usually becomes slightly worse with the addition of a tuned filter [3].
9. Using ERACS perform a H.I.S for the 3% and 10% cases to determine f_p and f_s and check if the tuning has shifted these points below the lowest harmonic generated by a six-pulse convertor, thus demonstrating the effect of the tuned filter circuit.
10. Determine the rating of the reactor. The reactor is rated at 110% of the fundamental current which it conducts. Filter reactors are usually air core. [9]

5. CASE STUDY

The following demonstrates the step-by-step methodology applied to the case study in question.

STEP 1: Tables 1 to 6 give details of some of the typical parameters used. Run LF to check if modelling is correct. A successful LF was obtained.

Table 1: Line and cable data

Branch Identifier	Cable1	Cable2	Ind line
Line / Cable rating	0.295kA	0.6kA	0.6kA
Nom. Rated Voltage	11kV	3.3kV	3.3kV
Conversion ID.	$\Omega/\Omega/\mu\text{f}$	$\Omega/\Omega/\mu$	$\Omega/\Omega/\mu\text{f}$
+ / - Seq. Series R	0.1513	0.1	0.01
+ / - Seq. Series X	0.084	0	0.604
+ / - Seq. Shunt B	0.461	0	0
Zero Seq. Series R	0.1513	0.1	0.01
Zero Seq. Series X	0.084	0	0.604
Zero Seq. Shunt B	0.461	0	0

Table 2: Motor data

MOTOR DATA			
Identifier	Library. Key	MVA	Rated kV
MOT1	MBS2	3	3.3

Table 3: Transformer data

TRANSFORMER DATA				
Identifier	MVA	kV	+ R%	+X%
T1	3.6	11/3.45	0.1	5.0

A 5% tap is used to increase the voltage to 3.45kV in order to overcome the volder in this case study.

Table 4: Shunt data

SHUNT DATA		
Shunt Identifier	Shunt 1	Cap 1
Shunt type	MW,MVAr	MW,MVA
Rated Voltage, KV	3.3	3.3
Unit Rating, MW	0.1	0
Unit rating, MVAR	0.075	-0.84

Table 5: Busbar data

BUSBAR DATA				
Busbar Identifier	SRC-V	Bus 1	Bus2	Bus3
Voltage, kV	11	11	3.3	3.3
Freq., Hz	50	50	50	50

Table 6: Generator data

GENERATOR DATA	
Generator Identifier	GEN-1
Generator type	Infinite
No. in group	1
Assigned Voltage, pu	1

STEP 2: The results of the H.I.S for the injection bus and bus 2 revealed no resonance point. See figure 2.

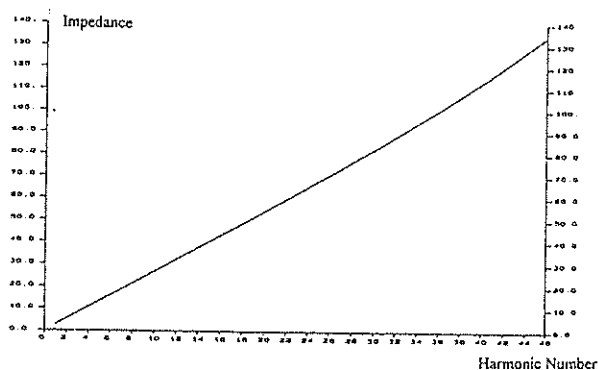


Figure 2: H.I.S without pf correction.

STEP 3: Perform a LF. The bus 2 LF flows are:

$$P = 2.579 \text{ MW} \quad Q = 1.594 \text{ MVAr}$$

From the P and Q values the pf was calculated to be 0.85.

The desired corrected pf is 0.96. To achieve this a Q value of 0.84 MVAr is required ($C = 224 \mu\text{F}$). The shunt modelling details are shown in table 4.

STEP 4: After the LF was re-run the desired pf value was obtained.

$$P = 2.576 \text{ MW} \quad Q = 0.767 \text{ MVAr}$$

$$\text{pf} = 0.96$$

STEP5: The H.I.S results are shown in figure 3. The resonant frequency point is the 7th harmonic.

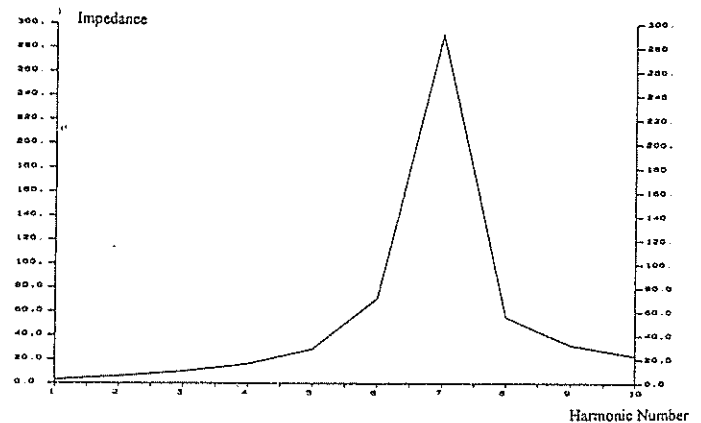


Figure 3: H.I.S with pf correction.

STEP 6: Calculate by hand the reactance value for the reactor. In order to shift the resonant point below the lowest harmonic generated by a six-pulse converter (5th harmonic) a reactor is connected in series with the capacitor bank to shift the resonance point to the 4.85th harmonic (3% below the 5th harmonic).

Using equations 1, 2, and 3 the following results were obtained:

$$f_s = 242.5 \text{ Hz} = 4.85^{\text{th}} \text{ harmonic}$$

$$L_F = 1.92 \text{ mH}$$

$$X_L = 0.604 \Omega$$

STEP 7: Model the tuned filter circuit. The reactor values in table 1 are for the 3% case.

STEP 8: The LF results showed the capacitor voltage to be 3.495kV. Pf capacitors are usually rated at 10% above nominal voltage. The capacitor voltage will thus not be exceeded. The pf has only slightly decreased to 0.95.

STEP 9: Perform a H.I.S. The result is shown in figure 4.

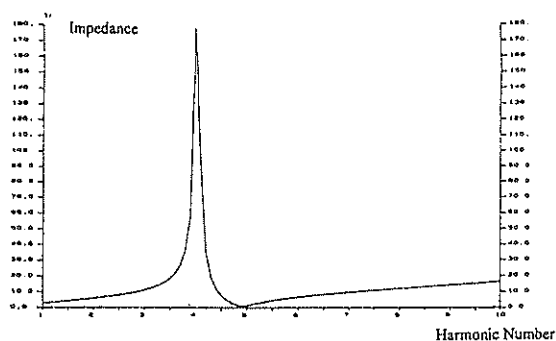


Figure 4: H.I.S. with 3% shift.

Steps 6, 7, 8 and 9 were repeated for a 10% filter. The values calculated for the 10% filter were as follows:

$$f_s = 225 \text{ Hz} = 4.5^{\text{th}} \text{ harmonic}$$

$$L_F = 2.23 \text{ mH}$$

$$X_L = 0.701 \Omega$$

The H.I.S. result is shown in figure 5.

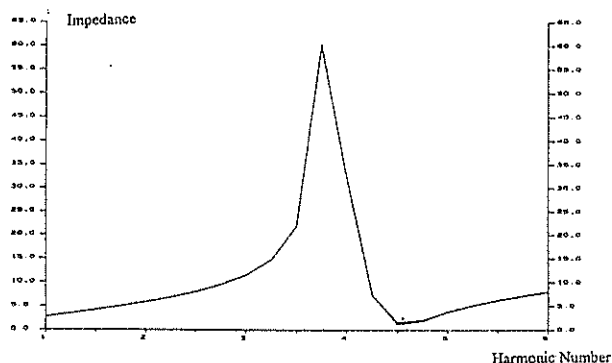


Figure 5: H.I.S. with 10% shift.

In both cases f_p and f_s were shifted below the 5th harmonic as desired.

Step 10: Determine the reactor rating. From LF (3% case) the fundamental current through the reactor is 0.139 kA. Therefore the reactor rating is $1.1 \times 0.139 \text{ kA} = 0.153 \text{ kA}$. For the 10% case the fundamental current is 0.138 kA giving a similar rating.

6. CONCLUSIONS

The methodology used in conjunction with the software package as an analytical tool was found to be successful as the desired results have been achieved. In harmonic resonance studies the X/R ratio of the transformer is important. It must be large otherwise f_p and f_s values do not shift below the lowest harmonic of the six-pulse convertor. The rating of the reactor is based on 110% of its fundamental current magnitude. Before pf correction

capacitors are installed a study should be performed to determine if resonance would occur at or near a harmonic frequency of a convertor. If so the capacitor bank has to be tuned away from this resonant point by adding a suitable reactor. The capacitor bank must be able to withstand the voltage rise across it due to the series reactor. This method could be used as an educational tool to teach the basics of power system harmonic resonance.

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CONTINGENCY ANALYSIS OF AN 11kV DISTRIBUTION NETWORK

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ABSTRACT

In this study a method of doing contingency analysis by means of computer software packages is proposed. Three packages are used to demonstrate comparative results. The study investigates the supply failure at a 66:11kV main substation where there is only one 66kV feeder. The area supplied by this main substation can however be fed via an 11kV backup supply, which is normally open. After a contingency, that is, a 66kV loss of supply, the alternative 11kV feed is closed. This paper investigates this load flow condition.

A further important aspect of this paper is the evaluation of the software packages and their suitability for contingency analysis.

1. INTRODUCTION

An electrical power network should be designed to continue operating, at acceptable levels, even when components of the network fail. Much emphasis is placed on the maintenance of the generation capacity and the EHV transmission network, but the 11kV distribution network is often neglected.

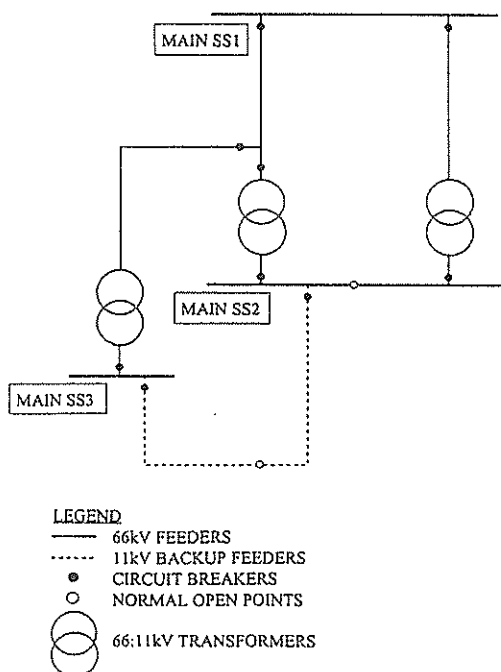


Figure 1: The 66kV Transmission Network

A "gut feel" approach is usually used and is totally reliant on the experience of the system operators responsible for maintaining the network. This is an unacceptable approach and a more formal procedure is required.

Figure 1 shows a simplified diagram of the 66kV network and backup 11kV feed which is used for the contingency study.

A contingency is defined as an event that causes one or more important components of a network, such as a cable, to fail.

A contingency analysis studies the effect of each contingency, in turn, on the remainder of the network.

This paper is only concerned with a load flow contingency study at the 11kV level.

2. CONTINGENCY ANALYSIS

A contingency analysis assesses the impact of a contingency or a set of contingencies on the state of the power system and identifies contingencies that would cause out-of-limit conditions.

Cable faults occur at unpredictable times and the faulted cable is taken out of service by the system protection operating. The remaining cables should be designed to maintain the system by taking the additional load immediately and still remain within their designed limits. The prolonged overloading of cables must be avoided.

Faults occur quickly and therefore, in most cases, operator intervention is not possible. A contingency analysis, using a load flow programme, can be used to set up the load flow for all contingencies in the system being studied.

Operators can then control the network defensively and develop operating constraints for the network. For each contingency studied a check on the loading of the remaining cables in the network is carried out.

A contingency analysis is also a vital tool used by system designers when additions and alterations to the network are required. An accurate analysis allows for sound engineering decisions to be made prior to the laying of new cables.

3. CASE STUDY

The contingency investigated was the loss of the 66kV supply to Main SS 3. This causes a supply loss to its 11kV substations. It was chosen because this is the "worst case" contingency possible on this network. Under such a contingency this area has to be fed via the 11kV backup feed, see Figure 1.

As a further verification control, the supply to Main SS 3 was disconnected in order to obtain actual load flow measurements on the 11kV backup feed and other cables. The results are shown in Table 4.

4. FIELD MEASUREMENTS

As computer simulations are to be used for contingency studies, it is necessary to firstly verify their accuracy. For this purpose field measurements are conducted which can be used to establish a reference platform.

The following field measurements were taken:

- Actual load flows on the cables during normal and abnormal network states so that these can be compared with the results obtained from the computer simulation contingency investigation.
- Cable resistance of cable C12, see Figure 2, which forms part of the 11kV backup feed, so that the actual cable resistance measurement could be compared with the value calculated using the cable manufacturer's data.
- Actual loads on the substations in the network for use in the computer simulations.

4.1 Load Flow Readings

The measurements at the Main Substations were obtained from the SCADA (Supervisory Control and Data Acquisition) system at the System Control Centre. All other measurements were taken using the ammeters at the Substations.

4.2 Cable Resistances

The resistance of cable C12, 6231 metres of 300mm² Al 11kV cable, from Main SS 2 to SS 8 was accurately measured. [1] The results are shown in Table 1.

Table 1: Field Measurements of Cable Resistance

Loop	Loop Resistance		Cable Resistance	
	DSM 200 Ductor Tester Ω	S-Test Set Ω	DSM200 Ductor Tester Ω	S-Test Set Ω
Red/White	1,360	1,326	0,680	0,663
Blue /Red	1,370	1,454	0,685	0,727
White/Blue	1,350	1,652	0,675	0,826

Note: The test was carried out with an air temperature of 18°C.

From the cable manufacturer's data the cable resistance can be calculated to be 0,748 Ω , at a soil temperature of 25°C.

The cable data supplied by the manufacturer confirms the measured reading of cable C12. From this it can be deduced that the manufacturer's values for the cables can be used with reasonable accuracy. Their values are thus used in this investigation.

4.3 Load Readings

Load readings at the substations in the network were obtained using the ammeters at the substations and samples of these readings are shown in Table 2.

Table 2: Load Readings

Substation	Voltage (KV)	Load	
		kVA	Amps
SS 5	11,66	210	11
SS 9	11,66	510	26
SS 11	11,66	1400	72
SS 12	11,66	2420	124
SS 16	11,66	250	13

5. SIMULATION

Computer-based simulation studies, using three software packages, were undertaken in order to determine their ability to predict real life contingency load flows. If the load flow results of this network verify the field measurements then computer modelling could be used for analysing any contingency in any 11kV distribution network.

The packages used were ERACS [2], SKM POWER*TOOLS [3] and NETAN3 [4].

These packages can all be used for power network analysis and are particularly useful for load flow studies.

A detailed one line diagram of the 11kV distribution network being analysed is shown in Figure 2. The network arrangement during normal conditions is shown.

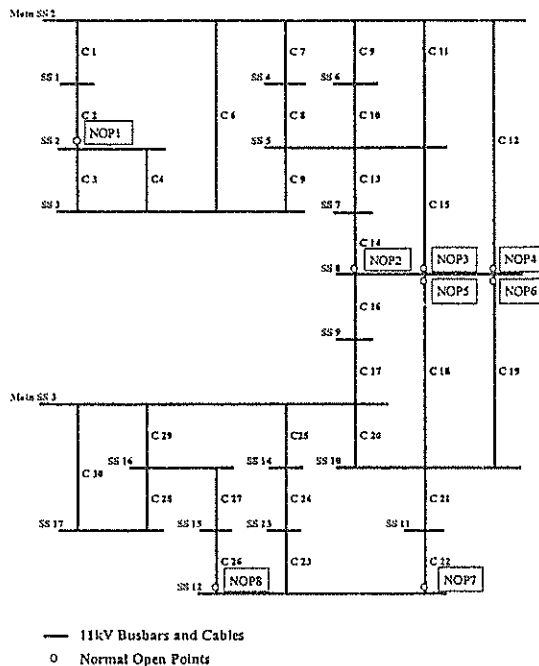


Figure 2: The 11kV Distribution Network

The 11kV back up feed is via cables C12, C14, C15, C16, C17, C18 and C19. When the contingency being studied occurs, normal open points NOP2, NOP3, NOP4, NOP5, NOP6 and NOP7 are closed. Closing NOP7 allows for a better load flow within the 11kV distribution network of Main SS 3.

Table 3: Cable modelling

Cable No.	Length (m)	Size of cable (mm ²) and Conductor Type	+*/- Sequence	
			Resistance (Ω)	Reactance (Ω)
C12	6231	300 Al	0,748	0,505
C17	1436	97 Al	0,488	0,131
C18	5656	120 Al	1,697	0,515
C19	3000	300 Al	0,360	0,243
C20	992	300 Al	0,119	0,080

Table 3 shows samples of how the network cables were modelled.

6. COMPARISON OF RESULTS

Table 4: Comparison of Load Flow Results

Cable No.	Field Measurements		Computer Simulated Load Flows		
	Before Contingency (Amps)	After Contingency (Amps)	ERACS (Amps)	Power* Tools (Amps)	NetAn3 (Amps)
C11	17	85	79	81	81
C12	0	216	229	237	233
C14	0	80	93	97	95
C15	0	120	110	119	116
C16	4	96	116	120	118
C17	31	73	90	94	92
C18	0	72	62	64	63
C19	0	240	254	262	257
C20	131	166	149	155	153
C25	175	146	140	145	143

A comparison of the results is shown in Table 4 and it can be seen that the simulated load flows are similar and, in most instances, compare favourably with the actual load flows.

The results obtained are sufficiently accurate for a contingency analysis of the network.

7. A PROPOSED METHOD FOR CONTINGENCY ANALYSIS

A flow chart, shown in Figure 3, proposes a method of carrying out a contingency analysis. The method is as follows:

- Create a computer model of the network in its normal operating state.
- Select a cable, C, to be out of service.
- Simulate a load flow for this contingency.
- Note any overloaded cables.
- Select the next cable, C+1, to be out of service.
- Simulate a load flow for this contingency.
- Continue until all contingencies have been analysed.

A plan for the operation of the network can be easily drawn up from the results obtained and can also include a list of all the contingencies in order of importance.

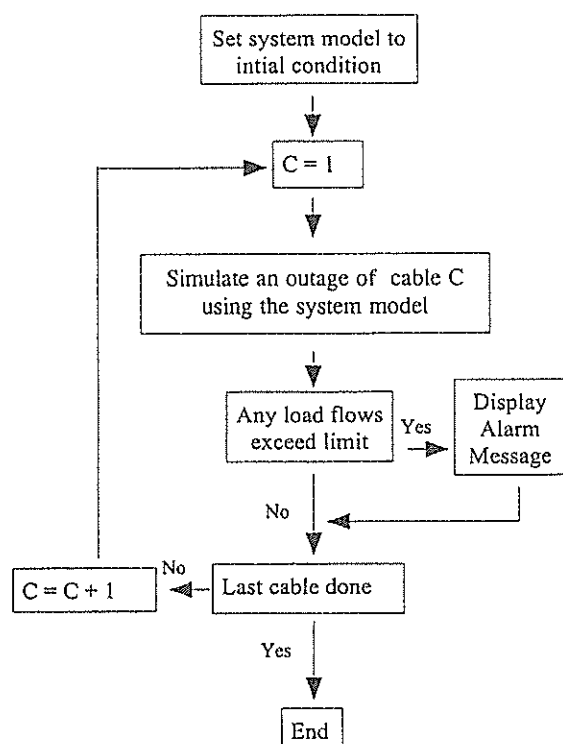


Figure 3: A Method of Contingency Analysis

8. CONCLUSION

- a. Contingency analysis is vital to the efficient and safe operation and maintenance of any 11kV distribution network.
- b. In order to provide an acceptable supply to the consumers it is foolhardy to rely on the experience of the system operators to maintain the system when a formal contingency analysis can provide an accurate and reliable method for operating an 11kV distribution network.
- c. The three computer packages all produced similar results and provide a simple and effective tool for the contingency analyst to use for network planning.
- d. The accuracy of the computer simulations is

totally dependent on the quality of the information used. The importance of accurate cable data cannot be over emphasised as incorrect information will greatly distort calculated load flows.

- e. It is recommended that where possible accurate field measurements of cable resistances be obtained and that prior to commissioning the resistances of all new cables are measured and recorded.

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ENERGY EFFICIENT LIGHTING FOR COMMERCIAL OFFICE BUILDINGS IN SOUTH AFRICA

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South African commercial buildings consume approximately 15,800 gigawatt-hours of electricity annually¹. Inefficient use of this electricity costs the nation millions of rands every year. The demand for this electricity is becoming greater, increasing the need for additional power stations. This is rapidly putting greater strain on the environment. There are certain energy efficiency measures and concepts that can be implemented in commercial buildings, that would create energy savings in cost effective ways. Recent developments in the lighting industry have resulted in highly energy efficient lighting equipment, which produce large energy savings and create long term investments in commercial buildings. These are described in this text.

Keywords: Energy efficient, cost effective, lighting equipment.

1. INTRODUCTION

According to research done in various office buildings throughout South Africa, lighting consumes on average 35% of the total electricity supplied to these buildings¹. This is a large proportion and thus has potential for energy savings. Air-conditioning systems consume about 50% of the supplied energy, but this text focuses only on the energy saving measures in lighting. It needs to be kept in mind though, that by reducing the lighting load, a large portion of a buildings heat load is reduced, and thus also the air-conditioning load. Energy efficiency regulations within the South African building industry do not currently exist, but would definitely ensure a reduction in the total energy consumption in many buildings, if they were to be implemented. Such regulations are currently being drawn up and proposed through the 'Green Buildings for Africa Programme'¹. However, according Prof. LJ Grobler, the project coordinator, they will probably only become official building regulations within about two years. In the mean time, energy efficiency measures in lighting and air-conditioning need to be researched, tested and implemented in South African commercial buildings, so that by the time the regulations are implemented, South African engineers already have the knowledge and understanding about the best, cost effective, energy saving design methods. A brief overview is given of how the energy consumption profile of a building's lighting equipment is obtained, and then ways of reducing this are described.

2. OBTAINING ENERGY CONSUMPTION PROFILES OF BUILDINGS: THE ENERGY AUDIT

Before any energy saving measures can be implemented in a building, a precise knowledge is required of where, when and how much energy is used. This is done by

what is known as an energy audit. An energy audit is a critical part of the energy savings process. Via the energy audit the engineer identifies a building's major energy users. The information gathered determines the energy saving measures that could possibly be implemented. It forms the 'base case' - like a fingerprint or footprint of the energy usage of the specific building. During 1997 and 1998 Eskom has been doing a number of energy audits in some of their own buildings. The drive of these energy audits has come from their 'Demand-Side-Management Programme', which intends to optimise and minimise South Africa's use of energy on the demand side². This programme has been implemented by Eskom to slow down the construction of additional power plants. The results of the energy audits done in these buildings intend to promote better ways in which energy management can be achieved, and that energy saving measures are not just energy saving initiatives, but long term investments.

A large part of the energy audit requires the engineer to determine exactly how much, when and where energy is consumed by the lighting equipment. This would involve the following steps:

- Recording the power rating, quantity and type of each lamp in each area of the building, measuring and recording the light levels in each of these areas, and questioning the occupants in each area on their opinion of the light levels. The type and amount of glazing also needs to be recorded for each area. This is done by walking through the building, while making observations and measurements, and recording these on as built drawings or tables drawn up for each floor.
- Finding out from the building manager at what times and by what means lights are switched on and

off in the various areas, over working days and weekends.

- Logging the electrical supply from which the lighting is tapped. This is used to verify the information gathered during the walk through lighting audit, described in the first point.
- Using the recorded observations to find areas in which inefficient lighting equipment is used, or where there is too much or too little light.
- More efficient lighting equipment, or any other means of saving energy need to be proposed. The measures taken cannot compromise the comfortable building environment that the occupants require.
- The cost effectiveness of these measures needs to be tested through cost benefit analyses and by determining the pay-back period of these measures.
- A report of the energy audit, with its findings of the most viable, cost effective energy reducing measures, needs to be compiled and presented to the building owners and manager, so that they can decide which of these measures should be implemented.

The following section describes some of the energy saving techniques that can be implemented within lighting.

3. REDUCING ENERGY CONSUMPTION WITHIN LIGHTING

3.1 Lamps

The lamp is that part of the lighting equipment in which light is created. Light can be created in various ways and thus there are different types of lamps. In the 1970's, and 80's most light in commercial buildings was produced by either incandescent lamps or standard fluorescent lamps. The conventional incandescent tungsten filament lamp converts about 96% of the total power it consumes into heat and only 4% into light⁵. This is because the tungsten filament requires to be heated to high temperatures before it emits useful visible light. Thus it is highly inefficient in converting useful and paid for energy into light, and also adds to the buildings heat load, and thus increases the energy consumption required for air-conditioning. With standard fluorescent lamps, the heat losses are only about 60% of the total input power. With the advances that have been made in the lighting industry, great improvements have been made in fluorescent lamps - they have been able to manufacture them in far smaller sizes (to the size of a normal 60W bulb), make them even more energy efficient, increase their rated life and also enable them to have a greater variety of colour appearances and colour rendering characteristics. The modern fluorescent lamp has split into two main new types: the *compact fluorescent lamp (CFL)*, and the

thinner fluorescent tubes. Both types have already had, and will continue to have a great impact on reducing commercial buildings' energy consumption and energy and maintenance costs.

a) Compact Fluorescent Lamps

The compact fluorescent lamp is a modern fluorescent lamp in which a number of smaller tubes are folded into the space of a 60W incandescent lamp. They can be screwed into the same fittings as incandescent lamps and can thus replace incandescent lamps. The compact fluorescent lamp consumes 80% less power than an incandescent lamp to produce the same amount of light. For this reason CFL's are rapidly replacing the function of the incandescent lamps and other lamps in most buildings around the world. A remarkable characteristic of the CFL is that it lasts up to 12 times longer than its equivalent. This means that considerable maintenance cost savings can be made by making fewer lamp replacements. All CFL's have a low power factor of about 0.5. This low power factor is not due to reactive components, but as a result of CFL's high harmonic content, and can therefore not be corrected by means of added capacitance. Good quality CFL's cost about 20 times as much as their incandescent equivalents, but in the long run they prove to be cost effective, through lower energy and maintenance costs. A cost comparison of an installation of 100 CFL's and 100 incandescent lamps, shows that annual electricity savings of R 9,800 and maintenance cost savings of R 2,000 can be made by the CFL installation⁸. The average pay-back period of a CFL installation is about 1 year. Thus such installations would create high return investments. However, their application is limited to certain functions because of their smaller size. While the lighting industry is still busy working out ways of reducing the harmonic content of CFL's, precautions need to be taken for certain installations where harmonics are undesired.

b) New Thinner Fluorescent Tubes⁶

For the larger open plan office areas and individual offices, fluorescent tubes still provide the best required average light level, with the fewest lamps. Technical improvements of fluorescent lamps have resulted in lamps which produce more light while using less power. The standard, older, fluorescent tube has a diameter of 38mm. The new fluorescent tubes have smaller diameters of 26mm (known as T26 or T8, the German and British metric systems respectively), and 16mm (known as T16). The standard fluorescent lamps have a luminous efficacy of about 80 lm/W. This means that the lamp puts out 80 lumens of light for every Watt of power supplied to the lamp. The new fluorescent lamps produce up to and between 93 and 106 lm/W⁴. They are thus up to 24% more efficient. Therefore considerable energy savings could be made by replacing all standard fluorescent tubes in a building with new tubes. The lamp

life of the new fluorescent lamp is 1,5 times longer. This will reduce the maintenance costs of the new lamps. The lamps can produce a better quality of colour and a greater variety of colours. This, together with the fact that tubes down to 7mm diameters are being produced, increases the functions and versatility of fluorescent light in the commercial building environment. The cost per lamp of a new T8 fluorescent tube is almost double that of a standard fluorescent. This means that initial costs of an installation with new fluorescent lamps should be higher. Often this is not the case, because less of the new lamps, which are brighter, are required for the same area. Therefore less lights and light fittings are required. Costs can thus be recovered, and additional savings can be made, through electricity savings and lower lamp replacement costs. A cost comparison done for a possible installation of fluorescent lamps in a South African building, to create an average light level of 500 lux, showed that almost R70,000 could be saved through installing less lamps and ballasts, using T8 fluorescent lamps, rather than standard fluorescent lamps³. The T8 installation showed that annual electricity cost savings of R40,000 and maintenance cost savings of R15,000 could be made. Such an installation would thus also create a high return investment, while improving the overall lighting conditions.

3.2 Ballasts

Nothing happens within a fluorescent lamp unless suitable control gear is provided. The control gear required consists of a starter and a ballast. The main function of these two, is to switch the electrodes and to limit the current through the lamp. There are two main types of control gear for fluorescent tubes, namely the standard ballast and the more modern electronic ballast.

a) Standard Ballasts⁴

The standard switch start ballast consists of an inductive resistor - a kind of rectifier. For a 50 Hz power supply these are large and heavy, and cause losses of up to about 30% of the supplied power. Losses are in the form of heat, which adds to the air-conditioning load.. They have a low uncorrected power factor which results in high reactive power, which pushes up the maximum demand and thus also the electricity costs. In standard control gear the starter and choke (inductor) are separate, making the control gear quite bulky. Standard ballasts cause lamps to flicker, creating eye-strain and headaches.

b) Electronic Ballasts⁴

New, electronic ballasts use electronic control gear for ignition and current limiting. The electronic devices are small and light, and impulses are at a much higher frequency. This results in higher lamp outputs, for less power consumed. Heat losses of electronic ballasts are

almost zero due to the light weight electronics, opposed to the bulky inductors and switches of standard ballasts. The lamp life of the electronically controlled fluorescent lamp increases by 67%. The electronic ballasts also have a much higher power factor, reducing the reactive load. The new electronic ballasts thus allow for considerable energy and maintenance cost savings to be made. Electronic ballasts do not cause any flicker. With the electronic ballasts the impulses are of a high frequency, beyond which the human eye cannot detect. Therefore switch-on and continuous light created by electronic ballasts is far more comfortable and causes no eye-strain or headaches. Fluorescent lamps are now also dimmable through certain modern electronic ballasts. These ballasts tend to increase the lamp lumen output by about 8%, and lamp life up to 210%, as a percentage of the basic electronic ballast. These ballasts can therefore be used as an energy and cost saving mechanism, depending on the application. The modern electronic ballasts are about 33% more expensive than the standard ballast. A cost comparison between an installation of 2200 standard ballasts and 700 electronic ballasts for the same building, showed that annual electricity cost savings of about R50,000 and maintenance cost savings of R15,000 could be made by the electronic ballast installation, even though the initial installation costs would be R80,000 more. Because of the lower maintenance and energy costs, such installations create high return investments with pay-back periods of about 2 years, while improving the overall light quality.

3.3 Luminaires

Luminaires are the fittings in which lamps lie, which allow for light produced by lamps to be reflected in various desired ways. With them, the light produced can be focused to certain areas or spread across larger areas. Therefore the luminaires play a large role in actually distributing the lamp light - light which requires energy, and costs money. The colour, shape and amount of reflective surface of the luminaire is critical in determining how much of the light produced by the lamp is actually transferred to the desired area. Certain colours and materials which have been used for luminaires in the past, absorb almost half of the produced light. Often luminaires are painted with certain colours so that they would suit the office interior appearance. This is often wasteful. Generally the most efficient luminaires have metallic-silver or white reflecting surfaces. Because of South Africa's dry, hot climate there is a lot of dust in buildings. Dust on luminaires greatly reduces the amount of reflected light. Often luminaires in such buildings simply need to be cleaned at regular intervals (about once every six months). This can greatly improve the lighting conditions without spending large sums of money. The shape of the luminaire determines the direction of the reflected light. Where office lighting is

concerned, luminaires with rectangular reflecting surfaces are generally the most inefficient, as the light is reflected across the luminaire, resulting in inter-reflected light - light which is lost within the luminaire. Luminaires with parabolic shaped reflectors are the most efficient in distributing the light uniformly into a large area. Where there is more than one lamp in a luminaire, it is important that the reflecting sheet separates the individual lamps. If this is not the case, the light of the lamps is inter-reflected, and in this way about 50% of the light from each lamp is lost. The number of luminaires per area is also a very important factor to keep in mind when wanting to save energy and money. New double parabolic luminaires have a very high light output and thus less of them are required for an area.

4. CONCLUSION

While there are not yet any regulations that enforce energy efficiency measures to be implemented in commercial buildings in South Africa, there is sufficient evidence that such measures in the field of lighting will result in considerable long term energy and maintenance cost savings. Such savings can create high return investments with pay-back periods of between 1 and 2 years. Such installations should thus definitely be implemented in commercial buildings, even without there being any regulations enforcing them. Most of the energy saving measures do not compromise the quality of the building environment, but create an even better environment. Thus, if implemented, such measures would definitely be effective in reducing South Africa's high energy demand and also improve the environment within South African buildings. South African building engineers need to gain an understanding of these benefits, and start to practise and promote such energy efficient design techniques.

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IMPLEMENTING AN ENERGY MANAGEMENT PROGRAMME AT A TERTIARY INSTITUTION: A CASE STUDY

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Abstract: South African tertiary institutions have an ideal opportunity to initiate energy management programmes on their campuses. The benefits of an energy management programme include financial efficiency, improved training of students and the creation of energy awareness amongst students, staff and local communities.

Key Words: Energy Management Programme, Energy Policy, AfriCampus, University of Pretoria

1. INTRODUCTION

The management of energy tends not to feature very high on the list of concerns of many University and Technikon communities. Usually only a few members of the support personnel at a University or Technikon are truly informed about the costs of energy consumption and energy wastage. In Africa, very few countries, commercial companies and tertiary institutions have formal energy management programmes in operation. Examples of tertiary institutions with established programmes include the University of Vermont, University of Warwick and the University of Sunderland.

2. THE UNIVERSITY OF PRETORIA

The University of Pretoria is situated in Pretoria in the Province of Gauteng, South Africa. A total of 12 faculties form part of this dual medium Afrikaans and English institution which has 27 000 students and 5 600 staff members attached to it [1].

Where used, the University of Pretoria is meant to imply the geographical locations in Pretoria, South Africa that contribute towards the overall financial responsibility of the University resulting from the consumption of energy resources:

- Main Campus
- South Campus
- Onderstepoort Veterinary Campus
- Medical Campus
- Women's Hostels
- Men's Hostels
- Sports Grounds and Facilities
- Experimental Agriculture Farm
- University Student Village (commune styled housing)

The energy and natural resources currently available to the University of Pretoria for which the University is financially accountable are electricity, water and gas. The University is supplied with electricity from the

City Council of Pretoria and receives 4 main accounts each month for Main Campus (combined with Men's Hostels), South Campus, Onderstepoort Veterinary Campus and Medical Campus. Although separated by nearly 2 kilometres, the Men's Hostels are supplied and metered from the same municipal sub-station as that of Main Campus. For this reason, the municipality combines these two accounts into a single monthly account and as a result of diversity, the University has enjoyed savings amounting to R 126,431.03 for the period from January to September 1998. The Women's hostels are directly connected to the reticulation network on Main Campus. The sport facilities and experimental agricultural farm are directly connected to the network at the Men's Hostels. Other than these 4 main accounts, the University receives a total of approximately 350 monthly electricity accounts from the City Council of Pretoria for all the various facilities distributed through Pretoria.

The University community that is targeted through the introduction of an energy management programme is illustrated in figure 1.

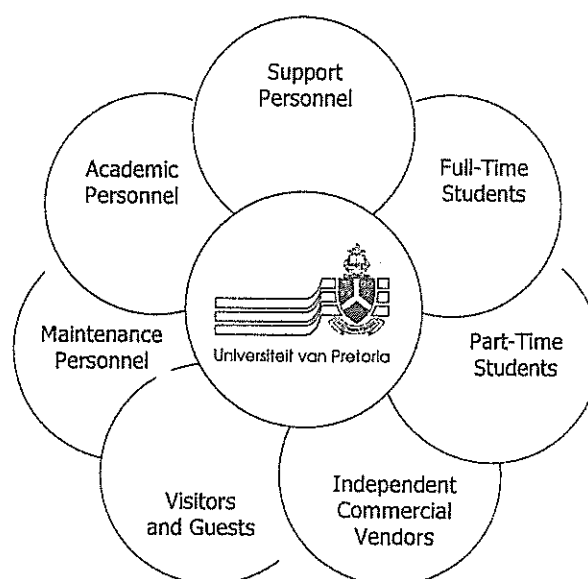


Figure 1: University Community

The staff members are classified into three categories, namely support personnel, academic personnel and maintenance personnel. The support personnel are those staff members that address the operational requirements of the University such as the financial and administrative tasks. The maintenance personnel are responsible for the general upkeep of the University premises and include duties such as gardening, cleaning and reticulation system maintenance.

Students are classified into two main groupings, namely part-time (referred to as Tuks-After-5 students) and full-time students. The student body at the University is the largest end-user group of the University and as such has a greater exposure to the end-use application of energy on campus.

The independent commercial vendors are those companies that operate on campus but are not directly part of the University of Pretoria. Examples include coffee shops, kiosks, fast food outlets and restaurants.

The last component of the University community involves visitors and guests to the University. Included here are those visitors who attend seminars and courses for non-degree purposes and various other functions.

3. THE NEED FOR AN ENERGY MANAGMENT PROGRAMME

The nature of the electrical load at the Main Campus of the University comprises two major elements namely a domestic component and a commercial component. These two main components are illustrated in figure 2.

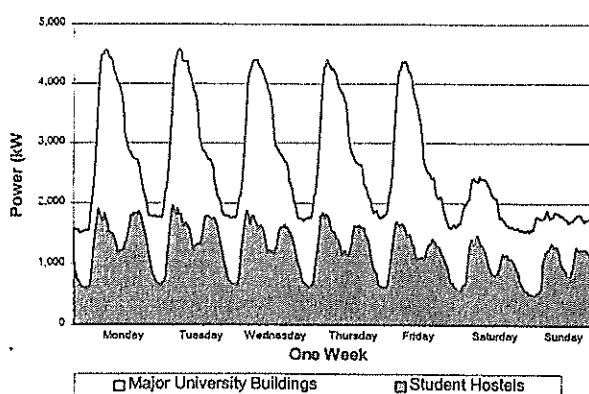


Figure 2: Commercial and Domestic components of the Main Campus Load (June 1998)

The profiles in figure 2 are taken from the week during which the Main Campus reached its maximum

electricity demand for the month of June 1998. The classic domestic profile present here is caused from the various student hostels that form part of the University electricity network. The effect of students that leave the hostels and travel home for the weekends can be seen on this profile from Friday afternoon through to Sunday evening. A profile of the major University buildings produces a near-classic commercial profile with a few exceptions. Usually a commercial company operates during set business times only with the occasional number of workers working either before or after hours. At the University however, at the time when a normal commercial company shuts down, the part-time students (Tuks-After-5 students) begin attending lectures. The profile also shows the effect of services that operate on Saturdays such as the University library. These two factors provide the commercial profile with its characteristics.

These two components provide the University with the opportunity to act as a miniature local supply authority due to the diversity of its load. The total electricity load on Main Campus taken from the same week is illustrated in figure 3.

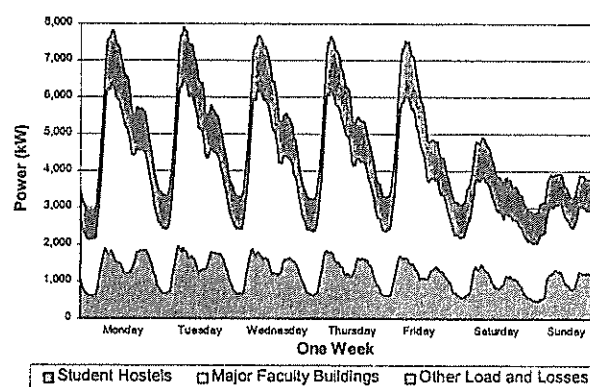


Figure 3: Disaggregated Load Profile of the Total Main Campus Load (June 1998)

The smaller faculty buildings as well as communal facilities that cannot be ascribed to any single faculty such as street and security lighting are grouped together to form the load labelled "other" in figure 3. The electric losses from cables and transformers are also included here. The University is currently billed for electricity according to a demand tariff. The demand tariff has two components namely an energy charge at a rate of 8.67 c/kWh and a demand charge at a rate of R44.50 per kVA (both exclude VAT) [2].

From figure 3 it can be seen that the maximum demand of the Main Campus of the University reaches just under 8 MW and that a base load of approximately 3 MW is ever present. The load factor for the total

consumption during this particular week was 61.53%, which is the present performance norm. At present, the average cost of electricity is approximately 19.77 cents per unit. This implies that the maximum demand charges contribute, on average, 11.1 cents per unit. The large base-load and the large maximum demand component are problems in terms of the type of tariff that is presently applied to the University. Through the implementation of various energy management projects as part of the energy policy, a great deal of attention can be focussed in altering the profile presented in figure 3 in order to suit the tariff structure imposed on the University.

4. PROGRAMME DIRECTION THROUGH AN ENERGY POLICY

The energy management programme obtains its direction through the adoption of an energy policy. According to definition, energy is considered as fuel and other resources used for the operation of machinery and policy is the course or general plan of action adopted by a government or party or person. An energy policy therefore implies a procedure or plan for achieving energy performance objectives.

The reasons for adopting an energy policy, as part of the energy management programme, include programme sustainability by ensuring that energy management becomes entrenched in the operating procedures of the University and programme transparency by ensuring that the rules and procedures relating to energy are applicable to all members of the University community.

During March of 1998, the Centre for New Electricity Studies formulated the draft Energy Policy of the University of Pretoria (EPUP). The EPUP is a general plan of action adopted by the University of Pretoria that highlights the commitment of the entire University community towards achieving a defined level of energy efficiency and management.

The primary emphasis of the EPUP is on the electrical energy resource. Although water is not applied as an energy resource at the University of Pretoria, it has been included in some of the areas of activity in order to control the wastage and utilisation of this precious resource. Gas is used on a small scale in specific laboratories and catering facilities. The cost of electricity is considered more substantial as both that of water and gas at the University.

The mission statement or purpose of the energy policy must be simple enough to easily invoke a measure of corporate commitment from the University community. It is felt that unless the members of the

University community cannot repeat verbatim the mission statement then it will not have the necessary impact. For this reason a simple mission statement was adopted as follows [3]:

"Manage the energy resources of the University of Pretoria to ensure maximum benefit to the University community with the minimum energy consumption and cost."

The above mission statement is addressed through the implementation of six areas of activity. These six areas should not be considered as stand alone options. In other words, they are not independent of each other and do not follow consecutively. Certain activities are vital towards more than one activity area. The reason for defining the six areas is an attempt to quantify the energy management programme in terms of a total quality management approach

Energy Measurement and Control

This involves the measurement of the electricity and water consumption according to the audit plan of the University. The control of a few key areas such as the hot water cylinders at the Women's and Men's hostels, the street and security lighting and the HVAC load of buildings is included here. The automatic control of "Smart" Lecture halls and the management and evaluation of new technologies through the implementation of various energy management projects are also targeted.

The Energy Management System (EMSup) presently enables the real-time measurement of the electricity consumption at all 14 sub-stations and major independent commercial vendors on Main Campus as well as the consumption of Main Campus and Men's Hostels from the Municipal Lynnwood Sub-Station.

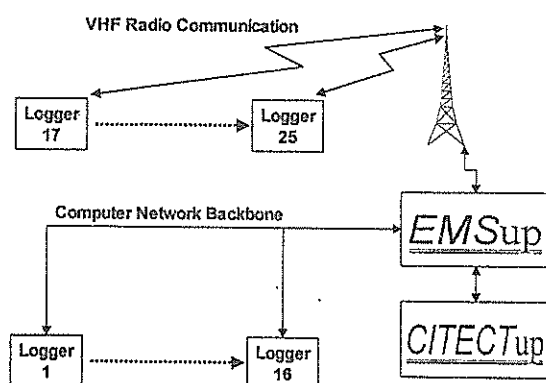


Figure 4: EMSup system components

The system communication requirements between the central data collector and the various data loggers is provided via the computer network backbone of the University. Where this network is not available, radio is used to collect data and transmit control signals.

The basic construction of the EMSup is illustrated in figure 4. The CITECTup component of the EMSup is a SCADA (supervisory, control and data acquisition) system used to verify the status of all the system components and to transmit the control signals to the various loads.

Energy Consumption Benchmarking

Benchmarks and specifications are vital towards following the performance of the EPUP. Typical benchmarks and specifications include monthly energy consumption benchmarks, lighting benchmarks, lighting specifications and various temperature specifications.

Energy Education

A comprehensive energy education campaign will be launched at the University that will aim to create an energy efficient culture amongst all the members of the University community. This campaign will keep the community constantly informed of the progress that the University is making towards energy efficiency.

Energy Economics

The area of energy economics comprises two facets namely, the control of the financial accountability of the University in terms of monthly accounts received from the City Council of Pretoria and the implementation of internal billing whereby end-user groups at the University are held financially accountable for the energy they consume.

Product Supply and Maintenance Contracts

Supply and maintenance contracts will be established to safeguard the interests of the University of Pretoria. Independent Contractors will have to ensure that their practices conform to the governmental safety regulations and all new constructions will need to conform to the specifications dictated in the EPUP before they are accepted. The supply of all technical documentation by Contractors will become compulsory.

Energy Marketing and Awareness

This specific objective involves the dynamic marketing of the EPUP itself to the University community, the University Alumni and the greater community as a whole. A heightened sense of awareness is paramount towards the success of the policy in terms of ownership by the University.

5. CONCLUSION

This paper highlights two major items. Firstly, it is plausible for a tertiary institution to implement and manage an energy management programme through

the adoption of an energy policy even though such an institution is very different from the commercial companies that have traditionally applied such policies. Secondly, that the accomplishment of any energy policy requires a high degree of community involvement that must be supported through managerial commitment. Undertaking a programme such as that at the University of Pretoria is vital from an African perspective since this type of policy strives to create an energy efficient culture which has tremendous social, economic and environmental impact.

Energy management is a skill that can be taught and learnt. People who become proficient in this skill within the University environment will begin to apply this knowledge within other spheres of their lives. In this manner, an efficiency culture is created. Although the energy management programme at the University of Pretoria is still in its infancy, the energy management team aims to reduce the present total annual cost of energy at the University of Pretoria during 1998 by at least 20% by the year 2002.

The Centre for New Electricity Studies has also initiated the AfriCampus collaboration between all tertiary educational institutions in Africa, an in particular Southern Africa. The purpose of this collaboration is to establish an energy management co-operation channel by which expertise, equipment and information can be transferred.

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OPTIMISED APPLICATION OF RENEWABLE ENERGY SOURCES IN RURAL ELECTRIFICATION

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Abstract:

Renewable energy sources are increasingly being applied in areas where grid extension is considered uneconomical. The initial costs of such systems are usually, but they can be minimised through proper sizing of the equipment, depending on levels of load demand and utilisation patterns. This paper discusses the development of a computer programme for determining the most cost effective energy source to supply the required load at any given time of the day.

Keywords: Rural, renewable energy, unit cost

1 INTRODUCTION

Renewable energy sources (RES) are increasingly being applied in electrification systems mainly because they are abundant and freely available, virtually pollution free and have comparatively low maintenance costs. However, this energy has not been exploited fully mainly due to the relatively high costs associated with the energy conversion technologies. In South Africa, many of the rural areas lying remotely from the grid, have a high potential of renewable energy, especially solar and wind. Rural loads are characterised by low load factor, which has a negative influence on plant operating costs, and makes it less cost effective to supply them from the grid. In such cases, renewable energy sources become the best alternative despite having comparatively high installation costs. The costs can be minimised if there is optimal load sizing by which the source and load characteristics are effectively matched. It is, therefore, essential to have accurate data on load demand and potential of RE source(s) of an area, before deciding on the type of source for a particular application. For the purpose

of this study the three types of RE sources considered were PV solar, wind and mini-hydro.

2 METHODOLOGY

2.1 Study area

The pilot area for the study was Maphephethe in the Valley of a Thousand Hills, about 40km west of Durban. The area has a hilly terrain with scattered households. At the centre of the village there is a high school, a health facility and a community centre consisting of a court room and offices of the area's Traditional Leader. The area has been considered by ESKOM to be remote and not economically viable for electrification by grid extension. It has adequate sunshine, low to moderate wind speeds and a few, mainly seasonal streams.

2.2 Data acquisition

A questionnaire was developed for the purpose of conducting a survey to determine the load demand. The obtained data was based on perceived demand because, apart from the high school, which has a PV solar system, no other household or building is

electrified. The level and pattern of demand were evaluated on the basis of appliances, which the customer would own, and their pattern of use. The data on potentials of the three types of renewable energy sources considered, was obtained from the meteorological and hydrological maps for the area. It was obtained as insolation levels, wind speeds as well as stream discharge rates and heads. Additional data was obtained from suppliers and users of RE systems for the purpose of creating a data base on unit costs.

2.3 Source evaluation

A TURBO PASCAL Ver 7.0 computer program was developed to determine the most cost effective RE source for given load conditions, taking into account both economic and technical factors. The potential output power and unit cost of each of the three sources were evaluated using the following equations:

$$P_{shp} = K * Q * H * \eta_t \quad (1)$$

Where : P_{shp} = Power from mini hydro in kW

Q = flow rate in m^3/s

η_t = turbine efficiency

H = effective head in m

K is constant including water density and the acceleration due to gravity

$$P_{pv} = G * A * \eta \quad (2)$$

Where: P_{pv} = power from solar in kW

G = solar irradiance in kW/m^2

A = panel area in m^2 , η = efficiency

$$P_{wtg} = A_r * k * V^3 * \eta_{ef} \quad (3)$$

Where: P_{wtg} = power from wind turbine generator in kW

A_r = area of rotor blades in m^2

V = wind speed in m/s

η_{ef} = effective efficiency

$$C_u = \frac{LCC}{Af * Cd * 365} \quad (4)$$

Where: C_u = unit cost in R/kW

A_f = annualisation factor = $\{((1+i)^n - 1) / i * (1+i)^n\}$

i = interest rate and n = number of years

LCC = life cycle cost in R

Cd = demand in kW

The unit cost was evaluated on the assumption that the load is for the whole year.

$$C_p = P_0 * C_u \quad (5)$$

Where: C_p = power cost in R

P_0 = unit output power in kW (as calculated using equations (1) to (3)).

The parameters required for the execution of the program were : time of day (for which the user wishes to evaluate power requirement), average daily wind speed (m/s), flow rate (m^3/s) and insolation level (kWh/m^2). Each of the sources was treated individually, that is, unavailability of one did not hinder the evaluation of the other sources. The program calculated the unit cost of each of the identified sources, identified the most cost effective source and the maximum power available from this source. In the modelling of PV solar sources, storage facilities (i.e. batteries) were also considered. The sizing of the battery was made to be an input parameter, which the user can specify.

The results of the program are stored in a separate file, which can be accessed any time. These results remain unchanged as long as the program is not executed again. So every time the program is executed, the old results are overwritten by the latest results. This feature is added to the program to allow the user to be able to study the results carefully. The

program can be adapted for application in any area. The flow chart of the program is shown in Figure 1, and a typical program output in Table 1.

3. DISCUSSION

The results shown in Table 1 are for specified input parameters, which can vary for individual customers, as well as, from area to area. The program has been developed to take care of such variations, and can therefore, be universally applied. Further work is being carried to improve the program so that it can also be used as an online power monitoring system for hybrid systems. By feeding in real time data on load demand, the program can determine the most cost effective source to supply the load. For areas where grid supply is an option, the program can be used to evaluate both the supplier and customer costs.

4. CONCLUSION

The developed program simplifies the task of determining the most cost effective renewable source to supply a given load, and is, therefore, a useful tool in systems load sizing. The reliability of the evaluated unit costs depends on the accuracy of data acquired on demand and potential of RE sources.

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Table 1: Sample program output

The program was executed at different loading conditions with other parameters fixed, i.e. flow rate, wind speed and irradiance.

annualisation factor is	:	8.51
time of evaluation is	:	14.00 Hours
flow rate is	:	3.00 m ³ /s
wind speed is	:	6.50 m/s
irradiance is	:	0.365 kWh/m ²

load is : 50.00 kW

Results are :

Power from shp is: 235.440 kW @ R 1.33 /kW
Power from wtg is: 94.560 kW @ R 0.85 /kW
Power from PV is: 4.184 kW @ R 2.15 /kW

Total potential is : 334.184 kW
Cost effective source output is 94.560 kW
Percentage of load suppliable by cheapest source is : 100

load is : 150.00 kW

Results are :

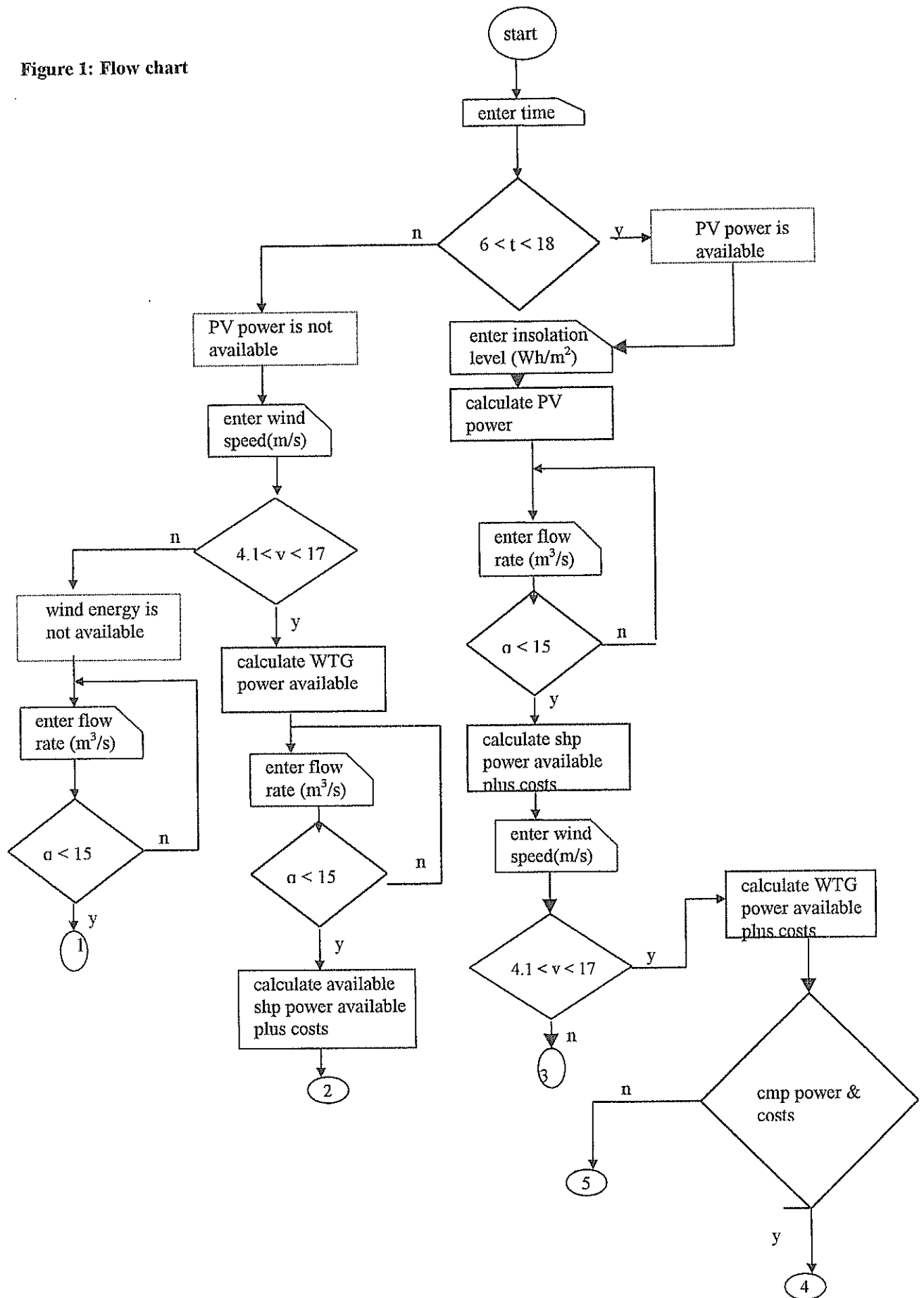
Power from shp is : 235.440 kW @ R0.44 /kW
Power from wtg is : 94.560 kW @ R0.28 /kW
Power from PV is : 4.184 kW @ R0.72 /kW

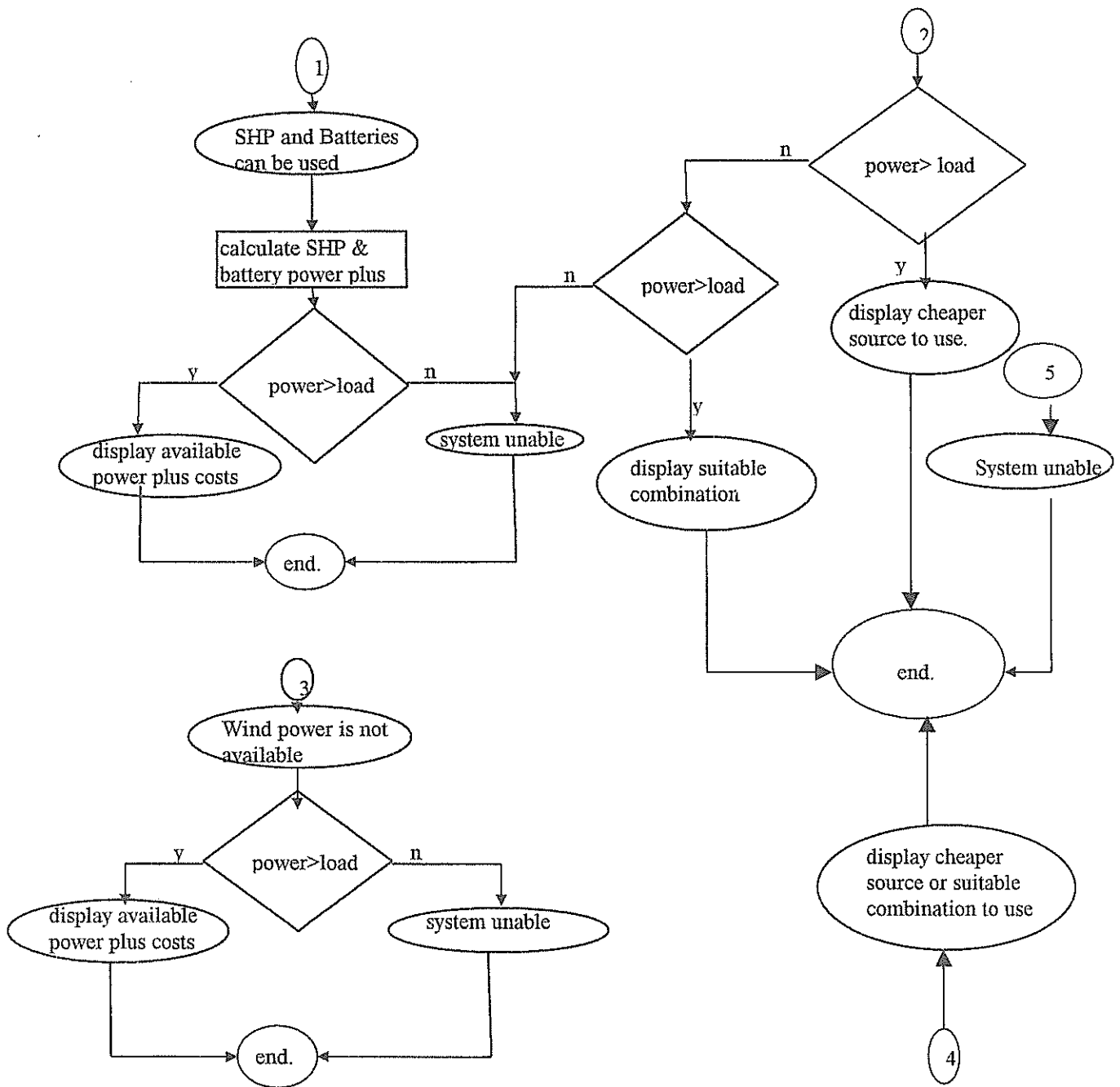
Total potential is : 334.184 kW
Cost effective source output is 94.560 kW
Percentage of load suppliable by cheapest source is: 63.040 %.
Balance obtainable from next available cheapest source.

ACKNOWLEDGEMENT

The authors wish to acknowledge with thanks the funding support received from ESKOM(TESP) and THRIP for this study.

Figure 1: Flow chart





CHARACTERISATION OF DEMAND AND LOAD FORECASTING IN NEWLY ELECTRIFIED AREAS

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Abstract: Load studies are being carried out in newly electrified areas to obtain more accurate data of their consumption patterns. The information is essential for cost effective designs of reticulation systems for such areas. This paper presents results of an artificial neural network model used to predict hourly peak demands of loads in a newly electrified area. Average forecast errors of less than 10% were achieved, using only limited historical data of the area's load patterns.

Keywords: load demand, neural networks, forecasting

1 INTRODUCTION

An increasing number of households in the informal settlements as well as rural areas are getting access to electricity, due to the wide spread electrification programmes undertaken by ESKOM and the local municipalities. Most of the consumers in these areas are low income earners, and their consumption is relatively low at poor load factors. Their daily demand curve is peaky in nature, with predominant morning and evening peaks[1]. Studies are being conducted nationally to understand the characteristics of demand for such areas and assess their impact on system load profiles.

Data obtained from such studies can be used in more accurate demand forecasting and optimised reticulation systems design. In the case of load growth, it becomes very important to identify the customer load pattern and predict the future electricity demand so that integrated system expansion planning, by considering the load

management alternatives, can be performed to meet the system peak demand more effectively. Determination of customer load characteristics provides utility companies better marketing strategies and also enables them to make full use of the power system apparatus for improved operation efficiency. In these studies load currents and voltages of different customers are recorded over 5 minute intervals using data loggers. The information is stored in a data base from which it can be accessed for various purposes. In this study, the data was used for generation of the daily load curves, and training of an artificial neural network(ANN), for short term load forecasting. The data analysed in this paper was obtained from Umlazi Township, about 25km south of Durban.

2 HISTORICAL DATA ANALYSIS

The load current of each customer was recorded at five minutes intervals. In order to have easy access to this data for analysis, a Turbo Pascal program was written to

compute the averaged half-hourly, hourly and daily load values, for a single or group of customers in each logger. Each logger could take up to seven customers. Once the averaged values for a particular customer were computed, they were then stored in a new file, and after all the customers' averaged values within a logger had been calculated, the respective files were merged using another program also in Turbo Pascal. The data in the merged files was then used to generate the load characteristics. It was also applied to the ANN model for load forecasting.

3 LOAD FORECASTING

3.1 Artificial Neural Networks (ANN)

In the past years, considerable effort has been made to develop effective approaches to forecasting the hourly loads for the next day or next week. The most commonly used methods are the statistical approach and the expert systems approach [2]. In this study, the latter approach has been considered, in the forecasting of hourly load demand, using artificial neural networks (ANN).

Neural networks are categorised by their architecture (number of layers), topology (feed-forward or recurrent) and learning regime. Most of the applications in power systems to date have used multi-layered feed-forward networks and use error back propagation learning [3].

The technique adjusts the connection weights w_{ji} of the signals passing through between neurons in order to produce an output, which minimises the total mean squared error of the network outputs, when compared with a desired output response pattern.

Artificial neural networks can be considered as a nonlinear statistical model identification technique, and

the architecture of most ANN models is designed by trial and error [3,4].

3.2 Forecasting model

Typically, the hourly load characteristics fall into two main and normally different patterns. That is, weekday and non-weekday demands. In the absence the load variation factors, such as the changes in the weather (especially temperature), the weekday pattern is preserved from Monday to Friday, and the non-weekday pattern applies to Saturday and Sunday. Therefore, in order to predict accurately the load pattern for a given day, historical load data is analysed by dividing all days into several groups, with each group comprising of days with similar load patterns. These days with similar patterns are referred to as the days of the same day type.

Once the day types are identified, then the forecast of the hourly loads for a particular day can be made. Several days in the past are chosen and then averaged to obtain hourly load patterns for these days. The hourly loads can be described by the vector

$$L_n = [L_n(1), L_n(2), \dots, L_n(24)]^T \quad (1)$$

where $L_n(i)$, $i = 1, 2, \dots, 24$, is the normalised load for hour i . $L_n(i)$ is related to the load $L(i)$ for hour i by the following equation

$$L_n(i) = \frac{L(i) - L_v}{L_p - L_v} \quad i = 1, 2, \dots, 24 \quad (2)$$

Hence the desired hourly loads $L(i)$ can be computed by using the equation

$$L(i) = L_v + (L_p - L_v)L_n(i) \quad (3)$$

Thus we still need an effective approach to forecast the daily peak load L_p and daily valley load L_v in order to get the hourly loads $L(i)$. It is important to note that the

daily peak and valley load are sensitive to weather variables such as temperature, humidity wind speed and cloud cover. The influence of temperature is more dominant, and therefore, the daily high and low temperatures are used as the weather variables in the short-term forecasting model.

3.3. Design and training of a multi-layer feed-forward neural network

Since the peak and valley loads for particular temperatures are known in advance, then a multi-layer feed-forward neural network is used. The neural network consists of the following layers:- the input layer, the output layer, and the hidden layer. For each neuron in the input layer, the neuron output is the same as the neuron input. For each neuron in the hidden layer, the net input is given by

$$\text{net}_j = \sum w_{ji} o_i \quad (4)$$

where o_i is the input pattern and w_{ji} is the connection weight from node i to j .

The neuron output is given by

$$o_j = \frac{1}{1 + \exp - (\text{net}_j + \theta_j)} \quad (5)$$

In order to generate the peak and valley loads an algorithm is used. The algorithm begins with assigning a set of random values for the connection weights. With a pattern p with desired output target vector

$t_p = [t_{p1}, t_{p2}, \dots, t_{pM}]^T$ presented, the connection weights can be updated using the equations:

$$\Delta w_{ji}(p) = \eta \varepsilon_{pj} o_{pi} + \alpha \Delta w_{ji}(p - 1) \quad (6)$$

where α is the momentum constant and η the learning rate, and

$$\varepsilon_{pj} = (t_{pj} - o_{pj}) o_{pj} (1 - o_{pj}) \quad (7)$$

These equations are based on minimising the error

$$E_p = 0.5 \sum (t_{pj} - o_{pj})^2 \quad (8)$$

It is to be noted that the learning rate η and momentum constant α in equation (6) affect the learning speed to a great extent. An algorithm was developed for evaluating the peak and valley values iteratively. The iterations are stopped when there is no significant change in the error function i.e. when $|E(I) - E(I-1)| < \beta$, where β is a small positive number. In the training of the network, the following must be considered:

Firstly, both the learning rate η and the momentum constant α affect the learning speed. It is recommended from [5] that a combination of $\eta = 0.25$ and $\alpha = 0.9$ could yield good results. Secondly, it is obvious from equation 5 that the output of the neural network must fall between (0, 1). In addition, the input signals should be kept small owing to the sigmoid function. Thus the input patterns must be first normalised. In the present work, the daily peak load and daily valley load (in A) are divided by 30 and the temperature (°C) is divided by 50 before they are used for training.

4.0 DISCUSSION OF RESULTS

Figures 1 and 2 show a comparison between the actual and forecast loads for the two day types (i.e. weekday and non-weekday). For the non-weekday, to obtain the hourly load pattern $Ln(i)$ of the day, the hourly load patterns of six previous weekends in April and May 1997 were fetched from the database and averaged. Also, the daily peak and valley loads for the six previous weekends were fetched, as well as the minimum and maximum temperatures of those days. These values

were then normalised and used as the inputs to the neural network. A similar procedure was followed for the weekday, but in this case, data for twelve previous week days was used.

An error analysis of the results showed that in the case of the weekend forecast, the model predicted the load very accurately between hours 1 and 8, as well as between hours 21 and 24. The average error in this case was about -6%, with a standard deviation of 4%.

In the case of the weekday load, the model was badly off in the first five hours of the day, but reasonably accurate for the rest of the day. The average error in the reasonably accurate range was about -8% at a standard deviation of 18%.

5 CONCLUSION

It has been demonstrated in this study that even with only a few days of hourly demand pattern data, a multilayer feedforward neural network can be used to predict the daily peak and valley loads reasonably accurately. This result is significant in newly electrified areas where load pattern data is scanty, and also whatever is available covers only a short period of time. The degree of accuracy in the forecast depends on the model assumptions as well as the accuracy of the input data. Further work is being carried out to improve the accuracy by varying the model parameters, which describe load changes in the different patterns.

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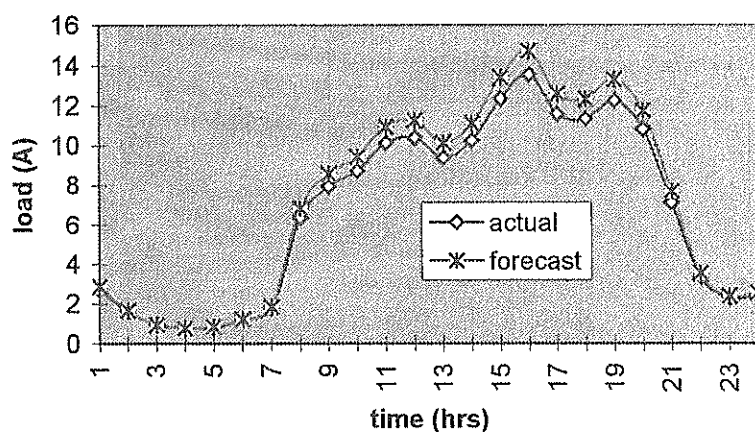


Figure 1: Comparison of actual and forecast loads on a week end day

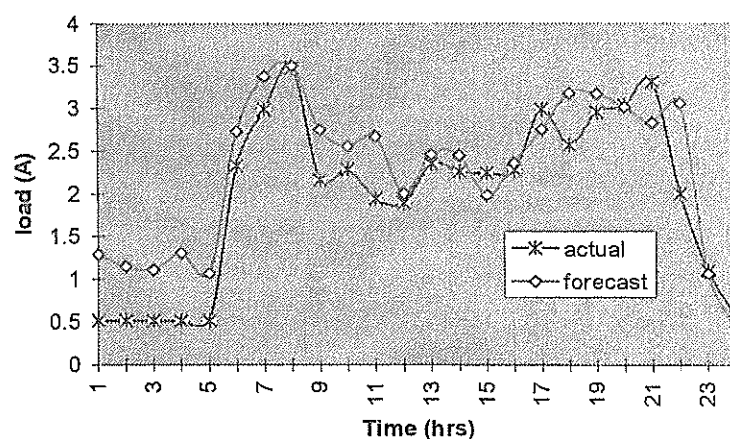


Figure 2: Comparison of actual and forecast loads
on
a week day



ELECTROMECHANICAL BATTERY FOR RURAL AREA POWER SUPPLY

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Abstract

This paper covers the technical aspects of an electromechanical battery. A permanent magnet motor generator is incorporated in a composite flywheel, running at high speed in a vacuum containment to minimise air friction losses. The flywheel is to be suspended on cascaded, Teflon coated, high-speed roller bearings. A composite flywheel requires a simple containment to sustain the vacuum for the moving parts, since most of the energy is dissipated in the fracture process in case of catastrophic failure. A 3-phase, switch mode bridge converter, being driven by a bipolar pulse width modulation board, achieves the variable speed control for the flywheel and the control of the DC bus voltage. The target storage capacity is 250 W-h and power rating of 40 W.

Keywords: Electromechanical, battery, flywheel

1. INTRODUCTION

Power supply to rural areas in South Africa many times takes the form of Photo voltaic cells connected to a lead-acid battery. The battery lifetime is about two years.

The availability of high strength fibre composites, low cost rare earth magnets and micro-electronics makes it possible to produce a non-polluting, cost effective battery.

This paper covers the main technical aspect of a stationary prototype electromechanical battery (EMB).

2. FLYWHEEL

2.1 SPECIFIC ENERGY

The specific energy for flywheel applications is Watt-hours of recoverable energy per kilogram of battery weight. This energy is proportional to the tensile strength of the material, volumetric- and shape-factors.

For a thin walled cylinder the specific energy is given by:

$$U = \frac{\sigma_{\max}}{2\rho} \dots\dots\dots (1)$$

U = specific energy in Joules/kg
 $\sigma_{\max}$ = maximum tensile strength
 ρ = density

Metal and short fibre composites can easily be formed into complex shapes. Filament wound

composites on the other hand is restricted to thin cylinders because of its orthotropic nature.

The optimal shape, which is a combination of an exponential and a thin disk, has an energy capacity of 1.9 times that of a cylinder.[1]
 Some typical values are shown in Table 1.

Material	$\sigma_{\max}$ [MPa]	U [W-h/kg]	U _{optimal} [W-h/kg]
Mild Steel	240	4.2	8.1
Aluminum	85	4.5	8.6
Mar. steel	1000	47	89.3
E-glass	760	73	139
Epoxy			
Graphite / Epoxy	4200	327	621
E-glass fiber	1700	190	-
Graphite fiber	7000	545	-

Table 1 Specific energies for some materials

2.2 FLYWHEEL PHYSICS

The Kinetic energy stored in a flywheel is given by:

$$P = \frac{1}{2} I \omega^2 \dots\dots\dots (2)$$

P = Kinetic Energy in Joules
 I = Mass moment of Inertia
 ω = Angular velocity in radians per second

The mass moment of Inertia for a cylinder is given by:

$$I = \frac{1}{2} m(r_o^2 + r_i^2) \dots\dots\dots (3)$$

m = Mass, and
 r_i, r_o = Cylinder inner and outer radii.

The hoop and radial stresses in a rotating cylinder are given by [8]:

$$\sigma_{\theta} = \frac{3+\nu}{4} \cdot \rho \cdot \omega^2 \cdot \left[(r_o)^2 + \frac{1-\nu}{3+\nu} \cdot (r_i)^2 \right] \dots (4)$$

and

$$\sigma_r = \frac{3+\nu}{8} \cdot \rho \cdot \omega^2 \cdot (r_o^2 - r_i^2) \dots (5)$$

σ_{θ} = maximum hoop stress

σ_r = maximum radial stress

ρ = density of the cylinder and

ν = Poisson's ratio

A fibre re-enforced plastic, filament wound, cylinder has a low tensile strength in the radial direction, equal or smaller than that of the matrix, and requires the use of a multiple rim design. The cylinders have to be connected with an elastic material that transfers angular and not radial forces between the rims.

For 250 W-h this yields a four rim composite E-glass fibre flywheel of the following dimensions.

Outer diameter = 200 mm

Length = 160 mm

Safe operating speed = 60000 rpm.

3 BEARINGS

3.1 MAGNETIC BEARINGS

Opposing ring and disk shaped NdFeB permanent magnets yields a stable radial bearing, but unstable (Earnshaw's Theorem [2]) in the axial direction. This is overcome by adding a mechanical, axial opposing disk magnet bearing pair or some active control. [2][3][7]

This configuration is self-aligning and allows for small unbalances in the flywheel.

3.2 CASCADED ROLLER BEARINGS

By cascading high-speed roller bearings, the speeds required can be achieved.

Precaution has to be taken when running in a vacuum, because the lubricant will evaporate. Lubricants exist, from the space program, which can work in absolute vacuum. For this prototype the bearings are Teflon coated as the means of lubrication.

The flywheel has to be highly balanced, limiting vibration.

A problem with cascading is that there is no means to ensure that the bearings share equal load.

4 CONTAINMENT

4.1 FRACTURE ENERGY DISSIPATION

Considering a 250 Wh flywheel, the power released is given by [2][5]:

$$W = \frac{P}{t} \dots (6)$$

W = Power in Watts

P = Energy in Joules

t = time in seconds

Time to dissipate energy (s)	Power [W]
10	25
5	50
1	250
0.1	25000

Table 2.1 Time-fracture Power

The duration of fracture thus greatly increases the power to be dissipated by the containment.

Using a composite flywheel eliminates the need for a massive and expensive containment, because fracture produces small, cotton wool like debris. The fracture process absorbs most of the energy, as opposed to a metal flywheel fracturing into large fragments.[2]

5. MOTOR/GENERATOR

5.1 PERMANENT MAGNET ROTOR

An approximation of a "Halbach array" [2] is used, constructed of rectangular magnets in a machined circular hub. This arrangement produces a uniform flux that is given by:

$$B = B_{rem} \log_e \left[\frac{r_2}{r_1} \right] \eta \dots (7)$$

where,

$$\eta = \frac{\sin \left(\frac{2\pi}{M} \right)}{\frac{2\pi}{M}} \dots (8)$$

B = Resultant uniform flux.

B_{rem} = Adjusted remnant flux in permanent magnets

r_2 = outer radius of Halbach array
 r_1 = inner radius of Halbach array
 M = number of poles

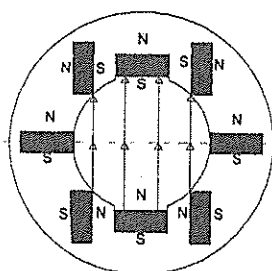


Fig 1: Approximated Halbach Array

A finite element analysis yielded a similar result and indicated that the theoretical formula is accurate enough for determining the flux density. Using 2x4x50mm NdFeB magnets in a 50mm diameter hub yields a uniform flux of 0.39 Tesla. This value has been confirmed by means of a Gauss meter.

5.2 THREE PHASE STATOR

The three phase coils consists of copper windings on a non-magnetic core. Since the permanent magnet field is always moving, eddy current losses will be generated in any magnetic and conductive material in the vicinity of the moving field.

The air gap between the coils and the rotor is only of value to the torque requirement as the flux is uniform in the Halbach array. Movement due to imbalances and heat expansion should be taken into consideration not to interfere with the windings.

In the prototype a stator diameter of 24 mm, 34 windings and a rotor inner diameter of 25 was used.

5.2 CAPACITY

The capacity is limited by the required charge / discharge rate, the size constraints and the cost. Since a low flux density, low ampere-turns and an iron-less core limit the power, the unit becomes only useful at substantial rotational speeds.

The equations[5]:

$$F = Bli, \dots\dots\dots (9)$$

$$T = Blir, \text{ and } \dots\dots\dots (10)$$

$$P = T\omega \dots\dots\dots (11)$$

F = Force
 B = Magnetic Flux
 L = length of conductor
 i = current
 r = radius
 T = Torque
 ω = angular velocity,

renders a power of 40 W at a speed of approximately 5000 rpm for the test system.

6 MOTOR DRIVE

6.1 CONTROLLER

The function of the controller is to determine the state of the flywheel, calculate the parameters and transfer them to the MA828 PWM board.

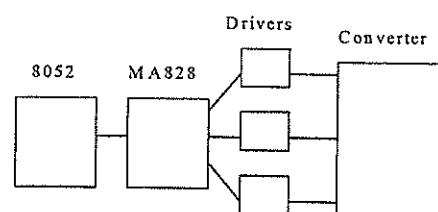


Fig 2: Control system

For the prototype an 8052 C-programmable micro controller is used in conjunction with a Marconi MA828 PWM controller and three Semikron SKHi 21 drivers.

6.2 PULSE WIDTH MODULATION

The 12-bit MA828 chip allows for accurate control of the motor generator in both the motoring and generating modes of operation.

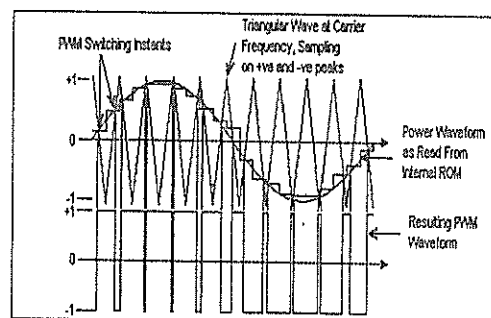


Fig 3: PWM switching scheme

This allows for the start up, where the windings have almost no resistance and a low power amplitude has to be used.

Using a bipolar switching scheme, at a carrier freq. Of 6 kHz, (see Fig 3) a sine wave is approximated.

Other possible configurations are unipolar switching, square wave operation or a Cyclo-converter [2][6]

6.3 SWITCH MODE BRIDGE CONVERTER

The switch mode converter is a frequency / voltage control device that allows for the flow of electricity into or out of the motor. The 120-degree shift between phases is automatically done by the MA828.

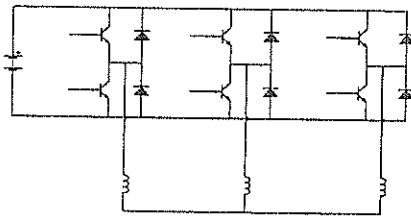


Fig 4: Three phase converter with load

Operating a switch-mode bridge converter, it is possible to create a cross conduction or "shoot-through" i.e. having both the top and bottom transistor switched on at the same time. The MA828 chip has a programmable pulse delay or blanking time to prevent this from occurring.[6]

In the linear region ($m_a \leq 1$), the fundamental frequency varies linearly with the amplitude component m_a

The line to line versus DC bus voltage relationship is given by [6]:

$$V_{LL} = \frac{\sqrt{3}}{2\sqrt{2}} m_a V_d \dots \dots \dots (12)$$

For $m_a \leq 1$

V_d = DC voltage

The high switching speed (6 kHz) requires that MOSFET or IGBT's be used, as to minimise switching losses and to be possible.

The motor AC voltage varies approximately linearly, with the speed, ranging from zero to about 500 Volts.

In rectification mode of operation the converter voltage is made to lag that of the motor, such that the DC voltage is constant.

7. CONCLUSION

The production cost of an EMB is far greater than that of conventional chemical one, but the service life makes it cost effective e.g. Typical lead-acid R300 for 2 year lifetime = R150 per year as opposed to R1000 for 10 years = R100 per year.

There is virtually no limit to the W-h capacity of the EMB.

Other possible stationary applications include power levelling, Uninterruptable power supplies and the like.

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IMPULSE FLASHOVER PERFORMANCE OF AN INSULATOR STRING AS A FUNCTION OF THE POSITION OF A BROKEN INSULATOR UNIT

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Abstract: This paper describes a study of the flashover behaviour of a 5 unit cap-and-pin insulator string as a function of the position of a broken unit. The configuration with the highest field intensity does not correlate with the setup having the lowest flashover value. It is concluded that the advancing leader dictates minimum flashover conditions.

1. INTRODUCTION

Live line workers are required to repair insulator strings with damaged units; to ensure safe working conditions the flashover performance needs to be known. Measurements have been made [1,2] which show that the position of the broken the broken units has a significant influence on the flashover voltage. No model of the actual breakdown process exists at present, however Rizk's work [3] with floating point objects is believed to have bearing on the flashover phenomenon. This study investigates the use of electric field analysis to predict the configuration with the lowest flashover voltage.

2. ELECTROSTATIC FIELD MODELLING

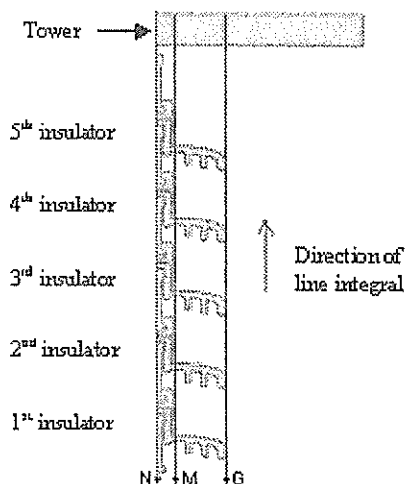


Figure 1: Insulator string assembly showing line integrals, with units numbered from HV end.

The insulator string was modelled using 2D rotational symmetry with all relevant dimensions suited to scale. The assembly includes a large rectangular body suspended above the grounded end in an attempt to account for the effects of the insulator tower.

Evaluation line integrals were placed at three different regions along the length of the insulator string. The first was extended along the radial axis of the string –

called the **Normal** position. The second line ran along the surface of the insulator caps –the **Metal** position, and the third line ran along the glass insulator edges – the **Glass position** (see Figure 1).

The line integrals were plotted for all positions of the broken insulator, including the full string, as a function of “field magnitude versus distance”.

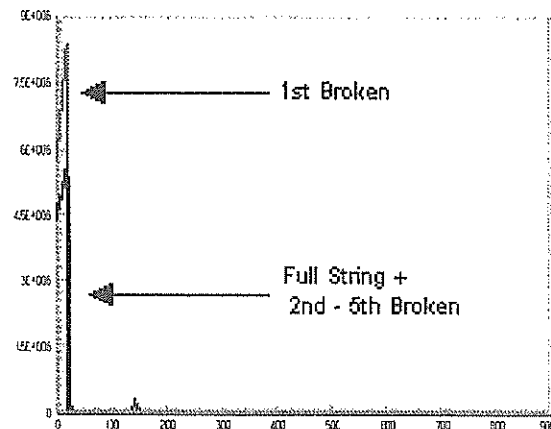


Figure 2: Comparison of E-field magnitudes for all cases (NORMAL).

Figure 2 indicates that the region along the string with the highest electric field is at the first insulator. This is not unexpected and is the consequence of the first insulator having the greatest voltage drop across it and hence the highest electric field. However, Normal indicates how the field magnitude at the 1st insulator is significantly higher when the 1st insulator itself is broken. The subsequent scenarios follow a similar trend, but of somewhat lower proportions.

The exact location of the highest electric field is as at the apex of the HV terminal, the pin extremity on insulator 1. Hence, the most likely place for the initiation of steamer breakdown is at the apex of the high voltage terminal.

The electric field study is not able to provide further insight beyond the identification of the most probable initiation point for flashover.

3. THE EXPERIMENT

A string of 5 Pilkington U70 glass insulators was used. Measurements were done on the complete string and with broken units in each of the 5 positions. The 50% flashover voltage was measured for 4 different rise times (56, 82, 120, 161 μ s) to determine the critical time-to-crest, and hence the minimum flashover voltage for each configuration.

The altitude of the test site is 1400m and for the duration of the test the ambient temperature varied from 22°C to 32°C. Each measurement was normalised to standard atmospheric conditions in accordance with IEC 60-1.

In the test set-up a metal rod was mounted at the HV end of the string to account for the effect of a high voltage conductor.

The tests were performed with positive switching impulses as negative impulses yield higher breakdown voltages. The results obtained are shown in Table 1.

Table 1: Results obtained from the 'Up-and-down' Method Tests performed.

Time to Crest \ Case Scenario	56 μ s	82 μ s	120 μ s	161 μ s
1 st Broken	437kV	440kV	418kV	430kV
2 nd Broken	460kV	449kV	436kV	445kV
3 rd Broken	451kV	435kV	407kV	429kV
4 th Broken	401kV	405kV	388kV	410kV
5 th Broken	371kV	381kV	368kV	377kV
Full String	462kV	455kV	441kV	444kV

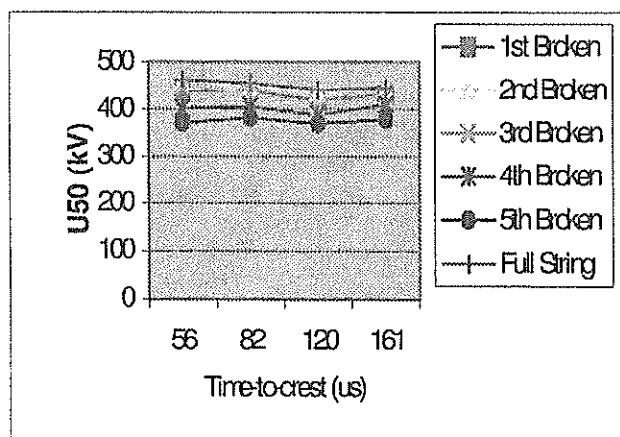


Figure 3: U50 versus time-to-crest for Total Comparison of Plots for all Case Scenarios.

From Figure 3 it can be seen that the configuration with the 5th insulator broken represents the lowest breakdown condition - some 17% less than that for the Full String.

4. CORRELATION OF RESULTS

4.1 The Contradiction

The electric field studies suggest that the configuration with the 1st insulator broken will yield the lowest flashover voltage. The measurements show that flashover is at a minimum for the broken unit in the 5th position.

The most likely place for steamer breakdown to ensue is at the apex of the high voltage terminal.

A positive leader grows as an extension of an ionised zone in the vicinity of the highly stressed electrode where a multitude of streamer-stems converge. The streamers themselves develop as a result of the corona activity in the region and hence the inception of the leader mechanism, which takes place after the formation of the streamer, will clearly occur at a voltage greater than or equal to the corona inception voltage.

The leader is essentially an extension of the electrode. As the leader breakdown begins and moves towards the grounded region, the entire dynamics of the electric field changes and an entirely new field arrangement evolves.

Hence the FEM allows for electric field analysis along the surface of the string prior to breakdown only. It does not permit any insight into the nature of the leader mechanism.

The field modelling for the evaluation of the electric field gives "steady state" conditions at the initiation of flashover. It cannot take cognisance of the development of the leader channel. Flashover is a dynamic process with the advancing leader changing field conditions as it lengthens.

4.2 The Breakdown Considerations

It is important to conceive how the arcing distance for all the scenarios is in fact not constant and as a result the length of the leader will vary. Figures 5a and 5b help illustrate this point by depicting the change of arcing lengths for the relevant string configurations.

It is assumed that the leader does not come into direct contact with the broken insulator on the string. Instead it follows an electric field gradient from the high voltage terminal through to the grounded terminal – taking the shortest possible route. The leader cannot in fact "see" ahead of itself, instead it follows a path as dictated by the streamer filaments in front of its tip.

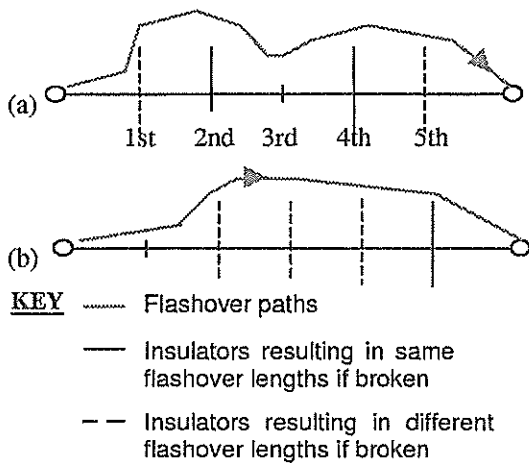


Figure 4: The change of arcing distance.

Referring to *Figure 4a*, it can be seen how the broken insulator in the 2nd, 3rd and 4th positions from the HV side effectively present the same arcing length and hence very similar leader components during breakdown. From *Figure 4b*, it is noticeable how for the broken insulator in the 1st and 5th positions, a similar arcing length is presented. This arcing length is definitely smaller than that of the three cases above. As a result, for breakdown of the string, the leader component is shorter for broken insulators at the 1st and 5th positions. The shorter arcing distance requires a shorter leader stem to bridge the gap and with a decrease in distance between the electrodes a lower voltage brings about breakdown.

Thus, the 1st and 5th positions of the broken unit both lower the insulation strength of the string to a greater degree than the three intermediate positions.

In determining the exact location for worst case, yet another factor is considered. The FEM simulations were run under the influence of a tower at the grounded end of the string. The effect of the tower served to moderate the electric field at the grounded end, and to some degree add uniformity to the field at its point of location. The tests were conducted without the inclusion of such a tower. Furthermore, the rod at the high voltage end of the string could in fact serve to replicate a potential tower at the energised region of the string.

The absence of a tower at the grounded end of the test setup would result in slightly higher, more inhomogeneous electric fields at this end of the string. Aggravating the situation further is the existence of a number of sharp objects at this end. These geometries will in turn serve to further enhance the electric field at the grounded end. Hence, what existed at the earth end of the insulator string during testing was a region of intense localised electric field. This situation will have a direct effect on the final stages of the leader, as it approaches the 5th and final insulator unit.

During its propagation, a streamer zone, the length of which increases slightly, during the leader propagation, precedes the leader tip. When the streamer zone reaches the grounded end of the string, the leader tip velocity increases exponentially, and "jumps" to the earth thereby completing the insulation breakdown.

If the E-field at the earth end is intense, the streamer ahead of the leader will be more developed and bridge the gap at some earlier period in time. As a result the "final jump" will occur at a lower inception voltage, and breakdown will ensue for a lower applied voltage.

At a broken insulator surface, due to the sharp edges, the E-field is more localised than for a whole insulator.

For the 5th insulator broken from the HV end, the E-field is more intense at this 5th broken unit than at the other insulator surfaces. Coupled with the effect of the electric field at the grounded end (due to the close proximity of the 5th insulator to the grounded end) the cumulative field will affect the continuous leader some distance away from the earth plane, and promote the inception of final jump further away from the grounded terminal and at a lower voltage.

As the position of the broken insulator moves further away from the grounded end, this coupling effect between the E-field at the ground region and the E-field at the surface of the broken insulator diminishes. Thus the total field intensity is lowered at the distances where the final jump inception is most likely to occur. As a result, the continuous leader must propagate closer to the high electric field at the grounded end as the broken insulator moves towards the HV end, in order to experience a field of high enough magnitude so as to initiate final jump.

For the length of string used in this project, it can be concluded that the flashover performance for the first two insulator positions are affected more so by initial conditions than by the final jump parameters.

4.3 Discussion

Hence, a theory is proposed - *The performance of an insulator string under switching impulse conditions is affected by three criteria:*

1. *The pre-breakdown electric field conditions.*
2. *The effective arcing distance of the configuration.*
3. *The position of the region along the flashover path in which the final jump is initiated.*

It is postulated that the condition of lowest required voltage for the complete flashover of the string, when a tower is situated at the ground end, exists for the broken insulator positioned closest to the ground end. Applying the above theory supports this statement and an explanation follows.

The pre-breakdown electric field conditions dictate the inception voltages of streamer activity. From these

streamers the development of leaders ensue. The propagation of these leaders are, however, no guarantee of complete breakdown of the string.

The FEM clearly indicated how the fields of highest magnitude prior to breakdown occurred for the 1st insulator closest to the HV terminal broken. However, as the leader extends towards the grounded end the entire dynamics of the field change and hence this position of the broken unit does not necessarily dictate the worst case scenario for the flashover of the string.

The existence of the shortest arcing distance for both the 1st and 5th insulators broken definitely lowers the breakdown voltage of the string for these scenarios. But the final jump criterion is deemed the most significant factor in affecting the required flashover voltages of the string, with relation to the position of the broken insulator.

A situation is envisaged for a conventional insulator string where continuous leader has been initiated...

For the position of the broken insulator closest to the grounded end, a localised electric field will exist at the surface of the broken insulator, due to its sharp edges.

As the leader propagates beyond the region of the 4th insulator, a highly distorted field is experienced which promotes the development of the streamer ahead of the leader.

Although the presence of the tower encourages the uniformity of the field at the region in question, the distortion of the field is unavoidable. The charge at the tip of the advancing leader itself has an electric field that modifies the surrounding environment. The localised field at the surface of the broken insulator couples with the field due to the presence of the leader, and results in a highly non-uniform field (*Figure 5*).

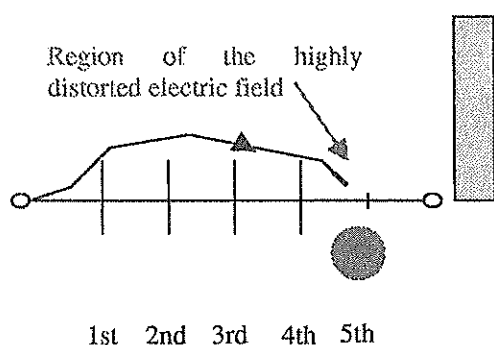


Figure 5: Representation of the region affecting final jump; for 5th insulator broken.

Thus bridging of the gap by the developed streamer occurs at some earlier period in time and as a result the final jump will occur at a lower inception voltage - and hence breakdown will ensue for a lower applied voltage.

For the 4th insulator from the high voltage end broken, as the leader propagates beyond the region of the 3rd insulator the distorted field is produced.

However, the advancing leader still has to contend with the 5th (undamaged) insulator unit. This 5th insulator tends to grade the electric field at the region of final jump and hence the streamer ahead of the leader, which ultimately bridges the gap, is somewhat less developed and the final jump occurs at a somewhat later period in time. Hence breakdown occurs at a higher voltage.

With increased distance of the broken insulator from the grounded end towards the HV terminal, the required breakdown voltage increases - but up until a certain point. This point depends upon the total length of the string. Hereafter the initial conditions predominate.

5. CONCLUSIONS

The configuration with the highest field intensities does not correlate with the configuration having the lowest flashover value and it is concluded that the advancing leader dictates minimum flashover conditions.

A theory is proposed showing how the flashover performance of an insulator string can be predicted with regards to three primary criteria:

1. The pre-breakdown electric field conditions.
2. The effective arcing distance of the configuration
3. The position of the region along the flashover path in which the final jump is initiated.

Applying these criteria the maximum reduction in the breakdown strength of the insulator string is found to occur for a broken unit at the position closest to the grounded end with an increase in insulation strength as the broken insulator migrates to the HV side.

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SURFACE AGEING MECHANISMS AND FAILURE OF SILICONE RUBBER INSULATION

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ABSTRACT

This paper identifies corona and local arcing as the major ageing influences on silicone rubber surfaces. The hydrophobic properties of silicone rubber lead to the formation of discrete water droplets on the surface and these can lead to local enhancement of the electric field and corona activity. This activity leads to surface ageing and a consequent loss of hydrophobicity. Once wetting takes place, surface currents flow and dry bands form. Arcing across dry bands also leads to ageing but this is not as deleterious as corona ageing.

KEYWORDS: Polymer insulators, corona, arcing

1. INTRODUCTION

Composite polymer (non-ceramic) insulators have been in use since the 1970s. Failures, in early installations, were related to [1]:

- a) Inadequate sealing against the ingress of moisture into the interface between the glass fibre core and its covering sheath;
- b) Acid attack on the core, leading to brittle-fracture (stress corrosion failure).
- c) Inappropriate choice of materials.

The manufacturers took steps to deal with these issues and, today, they are largely problems of the past. Polymer insulators, particularly those with high-temperature-vulcanised (HTV) silicone rubber as the core-sheath and shed material, are widely recognised as the preferred insulators in heavily polluted environments. This is largely because of the hydrophobic properties of silicone rubber and its ability to transfer this property to the pollution layer.

Despite the fact that the properties of polymers are vastly different from those of glass and porcelain, the design and selection criteria for polymer insulators are firmly based on the criteria applicable to glass and porcelain insulators (see IEC 815). The implicit assumption that has been made is that flashover will be the dominant failure mode in service and that the flashover mechanism is similar to that found in glass and porcelain insulators.

As far as silicone rubber insulators are concerned, experience has shown that, excluding bird related incidents, flashover is a very rare occurrence and that the most common cause of failure is mechanical. The sheath covering the resin-bonded glass fibre (RBGF)

core becomes damaged, exposing the core and leading to the possibility of brittle fracture.

Figure 1 shows an insulator taken from service, revealing clear evidence of ageing and puncture of the sheath material. Therefore, an understanding of ageing mechanisms is a prerequisite understanding failure mechanisms.

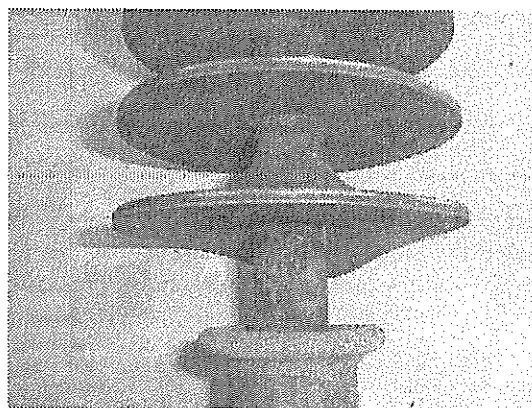


Figure 1: Service aged silicone rubber insulator.

2. CORONA AGEING

Recent studies have clearly shown permanent ageing effects of corona on silicone rubber [2],[3]. These are graphically illustrated in Figures 2 & 3. X-Ray Photoelectron Spectroscopy (XPS) has been used to determine the proportions of polymer and filler in the material, to a depth of 250nm, in virgin and corona-aged samples of HTV silicone rubber. Clearly, polymer has been removed from the rubber. Corona activity on the surface has made the polymer chains highly volatile by breaking them into short lengths. These short chains leave the surface and are lost to the insulator. Diffusion mechanisms within the material replenish the polymer on the surface and the process

continues. With time, the polymer within the body of the silicone rubber is depleted

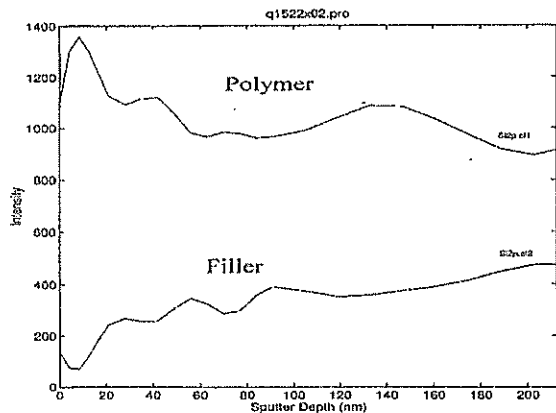


Figure 2: Depth profile for virgin, HTV silicone rubber [3].

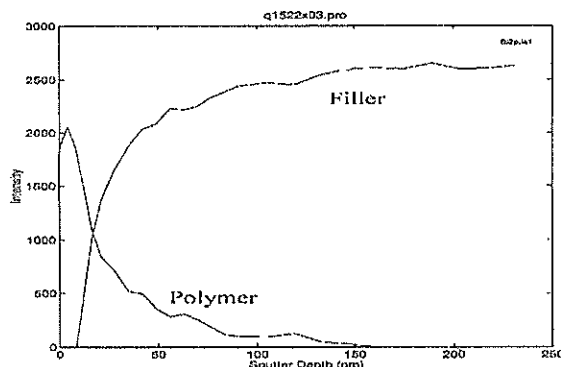


Figure 3: Silicone rubber sample corona-aged for 27 days and allowed 3.7 months for recovery[3].

and what polymer remains is found at the surface. Consequently, a severely aged sample of silicone rubber may exhibit significant hydrophobic properties, belying the seriousness of the ageing, which has taken place.

3. SOURCES OF CORONA

Insulators are designed to be free from corona in clean dry conditions. This is achieved by ensuring that the electric field is well below the corona threshold. The only way in which corona will develop on an insulator is if a field-enhancing factor is present. Recent investigations have revealed that water droplets act as very significant field enhancers for two reasons [4],[5]. Firstly, water has a very high relative permittivity of 80 and this causes a droplet to distort the field in the same manner as a conductor of the same shape. For a sphere, this is a factor of 3. The second feature of a water droplet in an electric field is that it distorts in the direction of the field, reducing the radius of curvature of the droplet and further enhancing the electric field.

Phillips et al [4] have shown that fields in the range 400-600V/mm, and above, give rise to corona on water droplets on silicone rubber surfaces.

The location of a water droplet needs careful consideration. If a droplet is located on the sheath, covering the RBGF core, the field enhancement will take place at the surface of the sheath. If corona is initiated it will attack the sheath. For droplets on the sheds, the field is more-or-less normal to the shed surface and the droplet distorts at its extremity, remote from the shed surface. Therefore, the influence of the corona on the silicone rubber is greatly diminished. Droplets located at the sheath-shed junctions will lead to corona attack on both shed and sheath. Observations made in the EPRI ageing chambers provide ample evidence of damage caused by water droplet corona [5].

4. DRY-BAND ARCING

In heavily polluted environments, characterised by extended wetting periods, silicone rubber loses its hydrophobic properties and a conducting film is formed along the insulator surface. This film carries a current from the high voltage connection to ground. Local heating, at positions of high current density and/or high surface resistivity, gives rise to dry bands, which break the current path and allow arcs to form. Figure 5 shows two silicone rubber composite insulators on test at the Koeberg Insulator Test Station (KIPTS) on the west coast of South Africa. When the photographs were taken both insulators had been subject to severe marine pollution for 15 months, energised as 19mm/kV.

Both insulators exhibited wetting along their full lengths and dry bands formed, giving rise to localised arcing. It is clear that the wet pollution forms a resistive path along the insulator surface that dictates the electric field distribution. The capacitive distribution, with the high field at the HV end of the insulator, present on dry insulators, is totally swamped by the resistive distribution. Dry bands can develop anywhere along the insulator length. This is the same phenomenon found on glass and porcelain insulators in polluted environments. The difference is that, in the case of the silicone rubber insulators, the magnitude of the leakage current pulses is approximately 10% of those found on glass and porcelain insulators in comparable situations. The arcs on the silicone rubber surface are extremely stable and will persist in the same position for days on end [6].

Experiments have been conducted to assess the ageing caused by dry band arcs on silicone rubber [6],[7].

While the work is not yet as comprehensive as that done for corona, it clearly indicates that dry-band arcing is far less deleterious to silicone rubber than

corona. This is illustrated in Figure 4 & Figure 5, where the hydrophobic performance of silicone rubber, exposed to both corona and dry band arcing, is compared. In these experiments, silicone rubber samples, taken from a naturally aged insulator, were exposed to typical levels of corona and arcing current. Figure [6] shows the influence of 8 hours of ageing on the contact angle as well as the recovery characteristics. It has been suggested that the difference in the wavelengths of the light emitted from corona and arcs account for the differences [7].

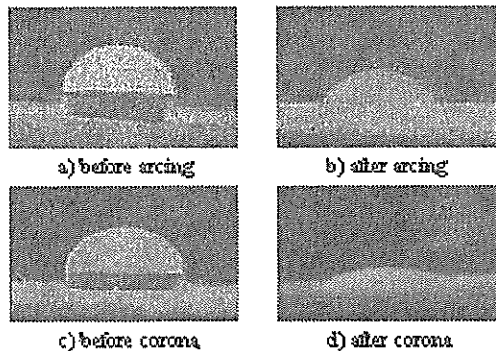


Figure 4: Contact angles for a 10 μ l drop of water before and after exposure to ageing [6].

After 15 months exposure a physical examination of the two insulators shown in Figure 5 did not reveal any apparent damage to the sheath or sheds at the locations of the arcs.

While more work still needs to be done, it is clear that dry bands arcing is likely to occur on silicone rubber insulators in polluted environments and that some ageing will result. Typically, the extent of this ageing appears to be small, but it should be minimised by restricting the arc current. This is achieved by maintaining an adequate specific creepage length, comparable to that recommended for glass and porcelain insulators. Values ranging from 20-35mm/kV, depending on the severity of the pollution, are suggested.

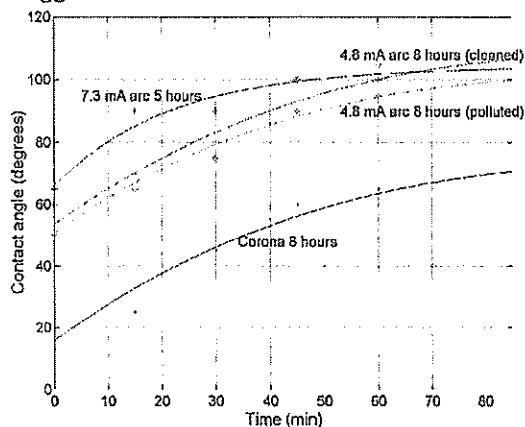


Figure 5: Recovery of hydrophobicity for HTV silicone rubber exposed to corona and typical dry-band arcing. [6]

5. STAGES IN THE AGEING PROCESS

1. When SR insulators are new, their surfaces are strongly hydrophobic, the only currents flowing through the insulator are capacitive [8]. The insulator has excellent performance [9].
2. With time, rain and wind cause pollution to adhere to the surface, increased roughness from UV radiation and embedded pollution aid pollution build up[10].
3. Low Molecular Weight (LMW) molecules diffuse onto the pollution, encapsulating it and making it hydrophobic.
4. Rain, fog, dew or high humidity form water droplets on the surface, even in areas where the droplets run off, tiny water droplets remain.
5. In areas where the field is high corona is emitted from water drops, causing a localised and temporary loss of hydrophobicity [5].
6. In parallel with corona ageing, the water interacts with the surface resulting in a reorientation of methyl groups, the "conformation change"[8], [2]. This results in a localised and temporary loss of hydrophobicity all over the insulator where water remains for sufficient time.
7. In areas where hydrophobicity is lost, the surface starts to wet, and dissolves some of the pollution forming a conductive layer. Some LMW's are dissolved in the water by the process of hydroxyl formation [8], further reducing hydrophobicity.
8. In addition, wet areas alter the field situation entirely, giving the possibility of water drop corona in hydrophobic areas where fields are too low under dry conditions.
9. The above processes continue until there is a continuous conductive layer over the surface. The formation of this layer is aided by the elongation of droplets by the electric field [5] and the hysteresis of the droplet shape. Water drop ribbons will form. Leakage currents start to increase. The thickness of the water film is much thinner than that on a porcelain insulator and leakage currents are much lower.
10. In areas of high current density and/or high resistivity, local drying of the conductive layer takes place and dry bands form.
11. Arcing and continued (more intense) corona causes further loss of hydrophobicity, erosion of the surface and tracking. Long term electrical activity however never seems to destroy all the reserves of LMW's. Electrical activity also breaks up longer polymer chains deeper in the material making them able to diffuse to the surface (LMW regeneration) and restore hydrophobicity when the ageing processes stop [11].
12. The above processes (5-11) will continue for the duration of the wet period. When the insulator dries, the ageing processes will cease and the

surface will start to recover hydrophobicity by the diffusion of LMW's to the surface and the reversal of the "conformation change".

13. When the next wet period begins, the ageing processes will resume, if the surface has fully recovered hydrophobicity, the ageing processes will start at steps 5 & 6. If recovery is incomplete, ageing will resume at a more advanced stage.
14. Full recovery is clearly a function of the degree of ageing and the degree of hydrophobicity before the wet period. However neither the type of rubber and nor treatment time have a significant effect on the initial speed of recovery [12].
15. The cycle will continue. The performance of the insulator will depend on the SR formulation, the insulator design and the stress grading. If the conductivity of the surface layer is very high the insulator may flashover, this should not severely affect the hydrophobic state of the insulator [7]. If the insulator has been badly designed, or there has been a manufacturing flaw) and there is insufficient SR on the sheath (particularly at the sheath-core interface, the erosion from dry band arcing may break through the SR and expose the fibre glass core which is susceptible to acid attack [5]. This is a very dangerous condition and may result in a line drop.

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A COMPARISON OF SIMULATION AND MEASUREMENT OF THE SHIELDING CHARACTERISTICS OF NON-FERROMAGNETIC CONDUIT CARRYING PARTIAL LIGHTNING CURRENTS

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Abstract: This paper investigates the shielding characteristics of a tubular, non-ferromagnetic conduit with partial lightning currents flowing along its length. Two computer models, the first based on a time domain solution to a lumped parameter approximation and the second based on the frequency dependant transfer impedance, are used to predict the nature of the induced interference voltages on cables lying within the shield. These predictions are verified with laboratory measurements.

Keywords – Partial Lightning Current, Skin Effect, Shielding, Conduit, Equipotentialisation, EMC, LEMP.

1. BACKGROUND

In South African industry, conduit and cable trays are very often not bonded at both ends. They are seldom thought of as shields for the reduction of electromagnetic interference (EMI) [1]. The latest lightning and EMC standards, both in South Africa (NRS 042 [2]) and internationally (SABS-IEC 61024-1 [3], 61024-1-1 [4], 61312-1 [5] and 61662 [6]) require such bonding to take place in an effort to create equipotential platforms which limit the formation of differential voltages between bonded areas or zones. The standards advocate the bonding together of all conduits, cable trays and shielding cabinets that should then also be bonded into the common earthing system [2, 5, 7].

In the event of a lightning strike, partial lightning currents flow along bonded shields with a minimum voltage build-up thus equipotentialising the site. However, these partial currents flowing on the outer surfaces of shields which house power and signal cables may give rise to significant interference voltages along the cables, particularly if the shield is long or has thin walls [1, 8]. A need exists, therefore, to quantify the effectiveness of cable shields under these conditions and then to emphasise their usefulness in reducing differential voltage magnitudes during lightning events.

2. INTRODUCTION

Two different computer models are considered in this paper. Firstly, a lumped parameter model that uses the concentric ring approximation proposed by Yen et al [9] was implemented in ATP [10]. The lumped parameter approximation caters for the skin effect under transient conditions and the model is thus capable of predicting the longitudinal electric field at the inner surface of the conduit. Secondly, a

frequency dependent transfer impedance-based model was implemented in MATLAB® [11]. This model incorporates the transfer impedance relation for cylindrical shields given by Vance [8] and derived originally from work by Schelkunoff [12].

To verify the results achieved by both models, a triaxial test fixture was constructed using a 3m length of aluminium conduit, having an external diameter of 19mm and wall thickness of 1,4mm. The triaxial fixture is well described in the literature [13, 14] and has been used for similar measurements by Kern et al [15]. Impulse currents were injected onto the shield and the resulting differential mode interference voltage waveforms are presented and compared to the predicted results.

3. TUBULAR SHIELDING CONCEPTS

For the purposes of this paper, the conduit with inner signal cables is treated as a coaxial arrangement with the disturbance current flowing on the outer shield only. Due to the skin effect, most of the partial lightning current will be forced to flow along the outer surface of the shield [1,15,16]. It is the magnitude of the current density at the inner surface of the tube, however, which determines the tangential (or axial) electric field (E_t), responsible for the induced interference voltages on the inner conductor.

$$E_t = \frac{J_t}{\sigma}$$

Where: E_t = Electric field vector tangential to the surface of the shield (V/m)

J_t = Current density vector in the axial direction i.e. tangential to the surface of the shield (A/m²).

σ = Conductivity of the shield material (Ω m).

4. TIME DOMAIN SOLUTION OF THE LUMPED PARAMETER MODEL

The time domain solution of the lumped parameter model is based on the partial resistance and partial inductance approximations suggested by Yen et al [9]. In this approximation, the tubular shield is divided into a number of concentric rings as shown in figure 1.

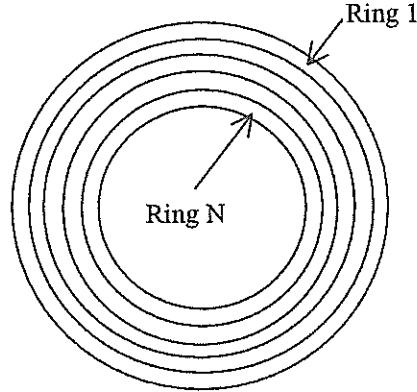


Figure 1: Tubular shield split into N concentric rings.

The ring resistances and the inter-ring inductances are given for the i th ring by [9]:

$$R_i = \frac{l}{A_i \sigma}$$

$$i = 1, 2, 3, \dots, N$$

$$L_i = \frac{\mu(r_{i-1} - r_i)}{2\pi r_i}$$

$$i = 2, 3, \dots, N$$

r_i = outer radius of the i th layer

A lumped circuit approximation may then be formulated as follows:

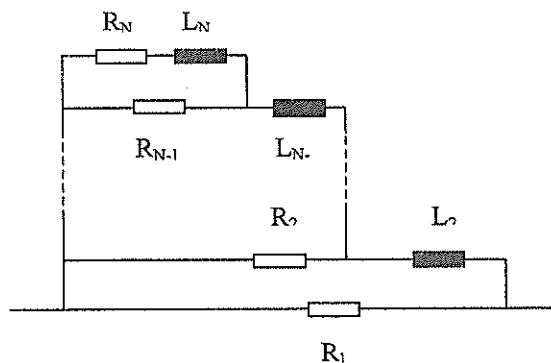


Figure 2: Lumped circuit approximation

The circuit shown in figure 2 was implemented in ATP as a lumped parameter model.

5. TRANSFER IMPEDANCE MODEL

The complex relation between the injected common mode current flowing on the tube and the electric field generated at the inner surface of the tube is given in general by the transfer impedance of the shield [8]:

$$dZ_t = \frac{E_t(\omega)}{I_{cm}(\omega)} dl$$

Where: dZ_t = Infinitesimal element of transfer impedance (Ω).

E_t = Tangential electric field (V/m).

I_{cm} = Common mode current flowing along the tube (A).

dl = Elemental axial length of the conduit (m)

By finding the spectrum of the injected current and multiplying by the transfer impedance term, the spectrum of the induced interference voltage, $V_{dm}(\omega)$ may be obtained.

$$V_{dm}(\omega) = Z_t \cdot I_{cm}(\omega)$$

In this paper, the discrete spectrum was calculated numerically using the FFT as it is not possible to calculate the continuous spectrum analytically due to the complex relationship between the transfer impedance and frequency.

Transforming the resulting spectrum back into the time domain gives the differential mode interference voltage waveform, $V_{dm}(t)$.

To determine the transfer impedance over the frequency range of interest for a tubular shield, we must first calculate the skin depths at these frequencies using the following equation [8]:

$$\delta = \frac{1}{\sqrt{\pi f \mu_0 \mu_r \sigma}}$$

Where: δ = Skin depth (m).

μ_0 = Permeability of free space (H/m).

μ_r = Relative permeability of the tube material (H/m).

f = Frequency (Hz)

σ = Conductivity (Ωm)

The transfer impedance for a solid walled tube is given by [8]:

$$Z_t = R_o \frac{kT}{\sinh(kT)}$$

Where: $R_o = 1/(2\pi\sigma T) =$ DC resistance of the shield per unit length(Ω).
 $k = (1+j)/\delta =$ The eddy current constant.
 $T =$ Thickness of the tube (m).

Figure 3 shows the predicted waveform for a 3m length of aluminium conduit with a 1kA, 8/20 μ s lightning impulse current flowing along its length. The conduit has an external diameter of 19mm and a wall thickness of 1.4mm. A comparative result from the lumped parameter model is plotted for the identical shield.

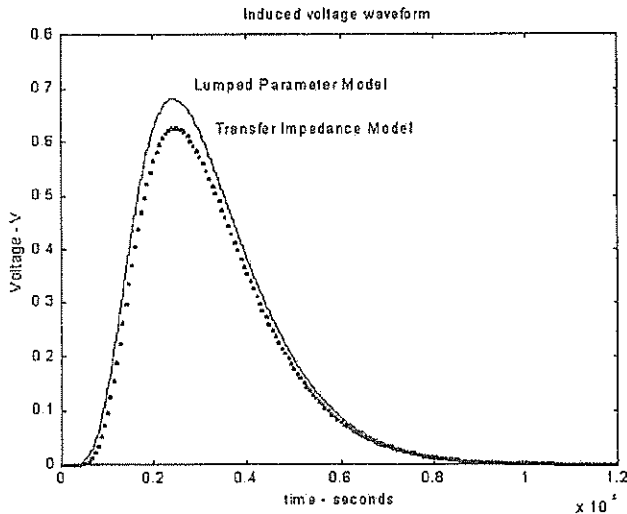


Figure 3: Predicted waveforms from the lumped parameter and frequency-dependent transfer impedance models.

6. EXPERIMENTAL PROCEDURE

Simulated results from both models were verified by laboratory measurements. As it is the tangential or longitudinal electric field at the inner surface of the conduit that leads to a differential voltage between the conduit and inner cables, it is necessary to take measurements that isolate the effects of this field. The most commonly used method of measuring the tangential electric field involves short circuiting the shield to the core at the current injection end and taking a measurement of the differential mode core voltage with respect to the shield at the other end of the tube.

Ideally, flux produced by the current return path should not link with the conduit as this leads to a further induced component due to magnetic coupling. For this reason, a triaxial test fixture was used [12, 13]. In the triaxial fixture, shown in figure 4, the shield and inner cables are placed coaxially inside a further tubular shield that acts as a return path for the injected current. In this way the linkage of flux from the current return path is minimised and the induced voltage which is measured is representative of the tangential electric field only.

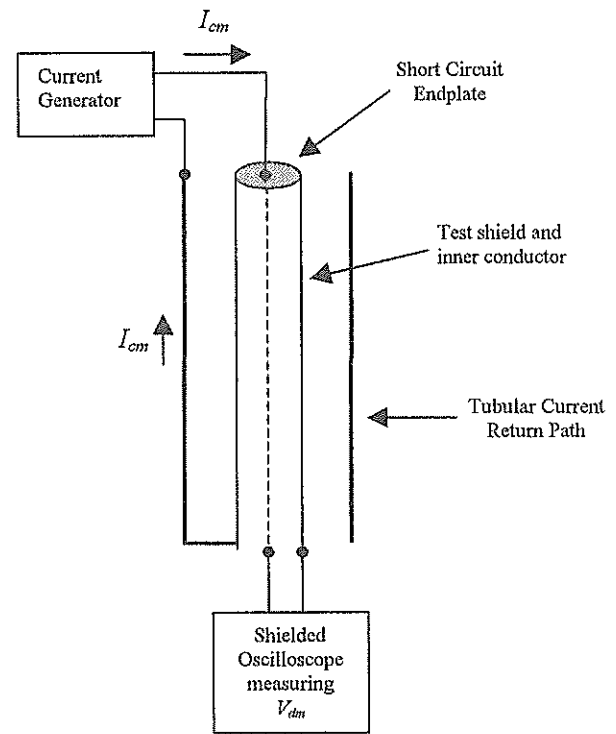


Figure 4: Diagram showing the triaxial arrangement and test setup. The triaxial setup eliminates magnetic coupling between the return path current to the shield under test and isolates the tangential electric field to be measured.

7. COMPARISON OF RESULTS

The test fixture described previously was used to inject an 8/20 μ s lightning impulse current with a peak value of 1kA onto an aluminium shield, 3m in length. The shield had an external diameter of 19mm and an average wall thickness of 1.4mm.

Using the measured waveforms and the predicted results from both models shown in figure 4, we may draw up a comparative table showing the key parameters:

Table 1: Comparison of peak voltage magnitudes and time delays to peak value for the predicted and measured waveforms.

Waveform	Peak Voltage (mV)	Time to peak (μ s)
Lumped parameter model	680	25
Frequency domain transfer impedance model	630	25
Measured waveform	700	24

Thus both models accurately predict the nature of the interference waveform. In this case, the lumped parameter model appears more accurate with a 3% difference in predicted peak magnitude (compared to a 10% difference for the transfer impedance model)

and 4% difference in predicted time to peak. The actual conduit under test, however, has a non-uniform wall thickness due to inaccuracies in the manufacturing process as shown below:

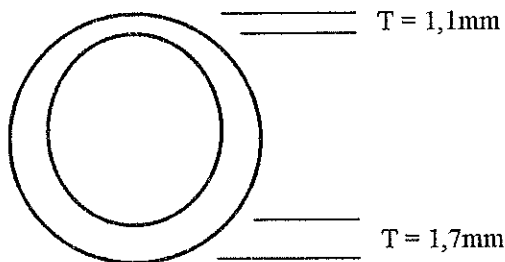


Figure 5: Diagram showing non-uniform wall thickness of the aluminium conduit used in the tests.

In both models an average wall thickness of 1,4mm was used, based on the measurements in *figure 5* which may not have been exactly representative of the average diameter of the shield. The measured results, however, indicate that it is a suitably good estimate to apply.

8. CONCLUSION

Lumped parameter and transfer impedance models that predict the induced differential mode interference voltages on signal cables lying inside a tubular non-ferromagnetic shield due to a common mode lightning impulse current flowing on the shield, have been presented.

Simulation results have been favourably compared to measured results and show good correlation in terms of both magnitude and time delay. Use of these models allows the accurate prediction of the nature of induced interference voltages for non-ferromagnetic conduit of varied diameter, wall thickness and conductivity. This is an important consideration in terms of the selection of surge protective devices for use on the electronic system inputs that have signal cables lying inside such shields.

Some preliminary simulations have also been performed on tubular ferromagnetic shields and the results show an improved shielding characteristic for steels with a high permeability. The models do not, however, cater for the non-linear behaviour of ferromagnetic materials under high current conditions at this stage.

9. ACKNOWLEDGEMENTS

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EVALUATION OF RISK IN PROBABILISTIC FEEDER DESIGN DUE TO STOCHASTIC LOAD CURRENTS

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Abstract: The latest LV distribution system design procedures advocates the use of the Herman Beta probabilistic approach to the analysis of the proposed network. This implies the choice and application of a level of confidence index or alternatively a level of risk. This paper attempts to evaluate the role that this choice has on the quality of supply to the consumer.

Keywords: Probabilistic methods, Herman Beta method, Risk, Distribution design

1. INTRODUCTION

Political changes in South Africa initiated an electrification drive and the national electricity utility, Eskom, made a commitment to connecting one and a half million households to the national grid from 1995 to 1999. The search for ways of reducing the cost per connection while still maintaining an acceptable quality of supply, produced various new technologies including new design algorithms and domestic load models.

The Herman Beta method [1] is a design algorithm for calculating the voltage drop in feeders supplying power to domestic consumers. The algorithm is probabilistic in nature and uses a parametric statistical description of the consumers load. Customer voltage levels are calculated and presented as a range of probable values. These voltages may be described in parametric form and a single design value may be extracted at specified level of confidence, or conversely, level of risk. In this context the risk has been defined as the design risk.

2. SOURCES OF UNCERTAINTY IN LV DESIGN

When designing LV feeders, the following sources of uncertainty may be present:

- The uncertainty in the prediction of the load parameters on which the design is based
- Sampling error in load measurement
- Inaccuracies in the statistical fit of the load model to the load data
- Inaccuracies of the network model
- The stochastic nature of the load currents

Each of these uncertainties introduces some probability of not being able to meet the primary design criterion: an acceptable quality of supply.

The investigation presented in this paper deals with the risk due to the stochastic nature of the load currents. Quality of supply (QOS) risk is defined as the risk of not meeting the design criterion due to the stochastic nature of the load currents.

The Herman Beta method uses a statistical description of the stochastic domestic loads at a specific moment in time to derive probabilistic voltage levels. The design risk mentioned earlier is therefore some indication of the risk in the design due to the stochastic nature of the load currents.

The design risk can be related to the QOS risk by taking into account the correlation between voltage levels at different time intervals.

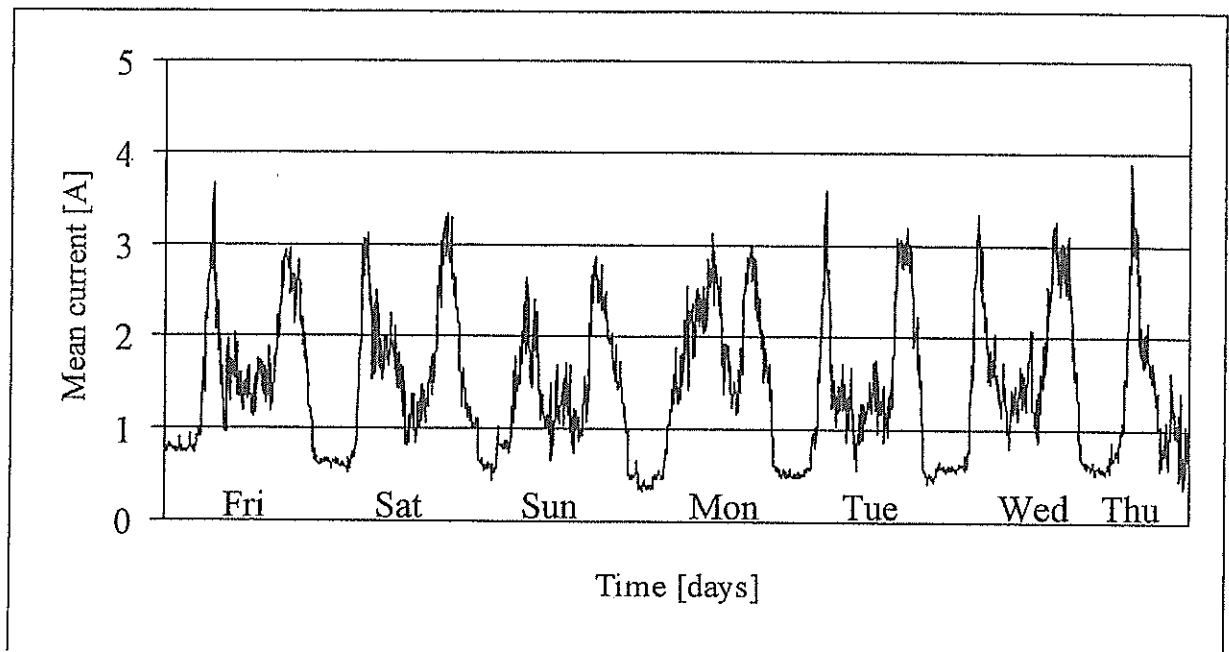
3. CORRELATION BETWEEN VOLTAGE LEVEL EVENTS

Consider the customer voltage level at any given time to be an identifiable event. Then in general terms, suppose that the probability of event A occurring is denoted by $P(A)$ and the probability of event B occurring is $P(B)$. It follows that the probability of both A and B occurring is $P(A \text{ and } B)$. Customer voltage levels are acceptable if they comply to statutory regulations.

If both events A and B were realizations of two acceptable voltage levels at different intervals, then $P(B)$ would be affected by the realization of event A. This is due to statistical dependence between the events. Correlation is a measure of how event A affects event B and is denoted by r .

Suppose event A occurs and $r = 1$ then $P(B) = 100\%$, however if $r = -1$ then $P(B) = 0\%$. The correlation coefficients between voltage level events were calculated using load data collected in Manyatseng. The correlation between voltage occurrences when the system was close to or at its maximum demand was between 0.6 and 0.8. The correlation is high enough to suggest that more than one voltage level occurrence should be considered.

Voltage level non performance events can occur at any instant but the probability of it occurring is greatest during certain intervals. These intervals coincide more-or-less with the periods of highest system demand. It is clear evident from the mean load profile for the week in



which the maximum demand occurs (figure 1), that there are several peaks of approximately the same magnitude. The QOS risk accounts for the probability due to all the possible events while the design risk accounts only for the probability of the event at the moment of the system maximum demand.

4. EVALUATION OF QOS RISK

The definition of an acceptable level of supply is required before the QOS risk can be evaluated. The QOS criterion for voltage performance suggested by NRS048 is the following:

The assessment period is a minimum of 7 continuous days.

On each phase of the supply voltage, for each 24h day (00:00 to 24:00), the highest 10 min root-mean-square values of the supply voltage within which the voltage remains for 95% of the time are recorded for each phase and the highest of these is retained as a daily value.

Similarly, the lowest 10 min root-mean-square values of the supply voltage within which the voltage remains for 95% of the time are recorded for each phase and the lowest of these is retained as a daily value.

The real QOS criterion is the consumer's perception of the voltage performance. The only way for the consumer to perceive violation of the QOS is through the failure of his appliances to perform adequately. A "once-off" occurrence of a QOS criterion violation is not likely to influence the consumer's perception of the quality of the supply he is receiving. Since this perception is unknown at present, an observability tolerance may be postulated as follows:

magnitude.

- A QOS violation only occurs if it happens on more than 4 days of the year.
- A QOS violation is considered to have taken place if the voltage performance is violated by more than 1%.

The QOS risk may now be evaluated using recorded consumer load data. The NRS Load Research Project [2] has collected a library of load data traces from various communities. These data are in the form of a series of load current readings with time stamps. Probabilistic load parameters (α , β and C_b) have also been calculated for each community.

The risk evaluation process simulates the voltage performance of a feeder using load traces from a particular township and then compares the results with the calculated performance.

Suppose a feeder has n number of connections then the simulation process for a specific feeder in a chosen community may be summarized as follows:

1. Randomly choose n consumers from the NRS library
2. Calculate the voltage performance for a specific feeder using the actual recorded load data
3. Evaluate the voltage performance using the QOS criterion defined above

4. Steps 1 to 3 is repeated N times and the number of times the QOS criterion is violated. The QOS risk for the feeder is calculated as:

$$\text{QOS risk} = N_v / N \quad \dots (1)$$

The choice of N is made to satisfy a predetermined level of confidence and the error may be calculated using the following equation:

$$\varepsilon = Z_{\frac{\alpha}{2}} \cdot \sqrt{\frac{N_v(1 - \frac{N_v}{N})}{N}} \quad \dots (2)$$

with
(1- α)100% is the required level of confidence
 $Z_{\alpha/2}$ is the quantile for the standard normal distribution

A convenient way of representing the relationship between the design risk and the QOS risk is the QOS risk/design risk ratio: QOR/dr.

5. SIMULATION RESULTS

Above mentioned calculations were performed on the load data of two townships: Manyatseng and Claremont. Manyatseng is a poor community situated in a rural environment and has been electrified for three years. Claremont is a medium income community in an urban environment and has been electrified for five years.

The simulations were done on several types of feeders and with different numbers of consumers. A design risk of 10% was used throughout.

In each case the following naming convention is employed to describe the connection configuration :

XXXmmm - n, where

XXX can be either

BAL - a balanced arrangement

COS - a cosine arrangement

mmm describes the grouping of consumers per node

n is the number of nodes on the feeder

Example:

The following three-phase, 4-wire, radial, LV, feeder is named COS400- 6. This implies the following customer connections:

Table 1: The arrangement associated with the name COS400- 6

Node	1	2	3	4	5	6
Red	4					4

is counted and stored as N_v

White		4			4	
Blue			4	4		

A phase resistance of 0.02 Ω is used in each test case and N = 1000 iterations are calculated for each.

Table 2: Results of simulation for Manyatseng

	qor/dr		
	n=3	n=6	n=9
BAL 111 - n	1.1	1.3	1.2
COS 300 - n	0.9	1.2	1.2
COS 400 - n	1.1	1.3	0.9

Table 3: Results of simulation for Claremont

	qor/dr		
	n=3	n=6	n=9
BAL 111 - n	1.6	1.3	1.0
COS 300 - n	1.2	1.6	1.0
COS 400 - n	1.0	1.3	0.8

The error in qor/dr calculated with (1) at a confidence of 95% is 0.15.

The qor/dr ratio varies for different connection strategies and for different number of consumers. It is further larger for Claremont than for Manyatseng.

The error incurred is rather large but due to the long computation time (almost two hours per test case on a Pentium 200MHz for 1000 iterations) it is considered adequate for inspecting general trends and sensitivities.

6. MEETING A QOS CRITERION WITH THE HERMAN BETA METHOD

If the qor/dr is known for a particular community and type of feeder, it can be used to modify the Herman Beta method to design a feeder for a QOS criterion.

The procedure may be summarized as follows:

1. Determine the qor/dr for a specific community and type of feeder

2. Specify a confidence level at which the QOS must be achieved (qor)
3. Calculate the design risk from the qor/dr ratio. The applicable confidence level is determined by the confidence at which the other uncertainties is dealt with and the influence the choice has on the lifetime cost of the system. For instance, the accuracy of the prediction tool used to determine the load parameters causes additional risk and must be absorbed in the QOS risk.

It is proposed that the qor/dr is calculated for various types of communities and ultimately a predictor is developed to estimate it.

The choice of QOS risk is currently being investigated and a guideline for choosing QOS risk is being formulated. The result will be used to steer the following:

- future decisions on network upgrades
- new networks
- accuracy of design tools
- accuracy of prediction tools

7. CONCLUSION

A method was developed to evaluate the risk due to the stochastic nature of domestic load currents. The results may be used to modify an existing probabilistic design method to design for a QOS criterion instead of a voltage level.

The study revealed some future areas of research which have already been undertaken and which will yield a management tool to ensure least cost designs. It will also be useful for evaluating the real performance of calculation procedures and load prediction algorithms.

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AUTOMATING THE ROUTING OF MEDIUM VOLTAGE ELECTRICAL CABLES IN ELECTRIFICATION DESIGN

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Abstract: This paper describes the method that will be used in developing a software tool to automate the routing of MV cable to interconnect a number of transformers on a CAD-based township map. The method proposes the use of intelligent map techniques to route the cables along roads and stand boundaries, while treating blocks of stands as impassible obstacles. This work forms part of the ongoing research at the University of the Witwatersrand to develop software tools to free electrification designers from as many repetitive and time-consuming tasks as possible.

Keywords: cable routing, intelligent map techniques, software tool for electrification design

1. Introduction

In its drive to bring power to the people, Eskom's target is to provide electricity to 1,320,000 households between the years 1998 and 2002. There is, however, great pressure on Eskom to reduce the cost per connection, because, according to current predictions, there will be a gap of R480 million/year between the actual and the target costs per connection [4].

The SED (Software for Electrical Distribution) group at the University of the Witwatersrand is developing a suite of software products aimed at reducing the cost per connection by liberating designers of electrification schemes from as many repetitive and time-consuming tasks as possible [1]. In work previously completed by the SED group, a software tool was developed which automatically divides a residential area into a number of transformer zones, using intelligent map techniques, and positions a transformer at the centre of each zone [2]. A natural progression from this work was to automate the routing of medium voltage electrical cables in a CAD environment, which is currently being researched and is presented in this paper.

At the 'New' Technologies for Electrification Seminar, held in Pretoria in March this year, papers by Eskom researchers and planners indicated that increasing use could be made of smaller transformers in future electrification projects.

This means that it will become increasingly important to be able to compare the costs of various configurations of MV feeder layouts connecting the transformers. It will also be important to be able to compare the cost of different MV technologies like SWER, dual phase and three-phase reticulation.

2. Review of existing routing algorithms

A number of algorithms have been developed for computer based routers and, initially, it seemed as if a router would be the obvious way to interconnect the transformers on a CAD drawing.

For transformers in a township, only binary case routers need be considered where the whole map is divided into areas that are either obstacles, where cable may not be routed, or free space where cables may be routed. All roads and stand boundaries, for example, would be available for the routing of cables, but cables would not be allowed to cross stands, so blocks of stands would be obstacles. It is not necessary to consider more complicated routers that can solve general weighted-region problems, which include cost ratings for the traversable areas. It was assumed that the cost differences between various regions of the same township would be negligible.

The three most commonly used binary case routers are maze routers, channel routers and linear expansion routers. A comparison between the three types of routers has shown that only the maze routers are guaranteed to find a path between two points, if one exists. It is also the only method guaranteed to find the shortest path between two points [3].

The most commonly referenced sequential maze-routing algorithm in literature is Lee's algorithm [5]. Lee's algorithm is, however, slow because it produces a wavefront, which expands from the source cell in all directions, along a grid, until it reaches the target cell. It becomes proportionally slower as the distance between two points increases. For two points N cells apart, the time required to find a path between them, is proportional to N^2 [3]. An automated MV cable router would be useful only if it were as fast as

or faster than manual routing methods, so maze routers do not seem to offer a viable solution for routing cables on maps of large townships with several transformers.

3. Finding the shortest route connecting a number of transformers

Instead of using a binary type router, the solution proposed in this paper is based on the assumption that the straight-line distance between two transformers is approximately a constant ratio compared to the length of the cable route connecting them. To find the most cost-effective way to interconnect a number of transformers, we have therefore assumed that those with the shortest straight-line distance between them, would also be closest together on a map.

For this work, we assume that we have a township map in a CAD format, which has been subdivided into transformer zones. A transformer has been positioned in each zone and the co-ordinates of all the transformers are available in a database, which is linked to the CAD file.

To find the most cost-effective way of interconnecting more than two transformers, the procedure would be as follows :

Step 1: Use the co-ordinates of the transformers to calculate the straight-line distance between each transformer and all the other transformers. Compile a list that records the straight-line distances and the names of both end-point transformers, sorted from the shortest to the longest distance.

Step 2: The straight-line distance list is now used to construct a network of transformers. The assumptions used for constructing the network are :

- Each transformer may be connected to no more than two other transformers;
- A test must be performed to ensure that all the transformers have been connected to the network.
-

The shortest distance is selected from the straight-line distance list and entered on a network list. On a transformer list, the two end-point transformers are flagged, to indicate that they have each been connected once. This process is repeated with the next shortest straight line. When a transformer has been flagged twice, all other distances in the straight-line distance list using that transformer become unavailable to the network list. Each time when a connection has been added to the network list, a

check must be done to establish whether all the transformers have been connected.

Step 3: In graph theory, algorithms have been developed to test the connectedness of a network. Once the checking in Step 2 above shows that each transformer has been used at least once in the network, a connectedness test must be performed, otherwise a situation as in Fig. 1 below could arise.

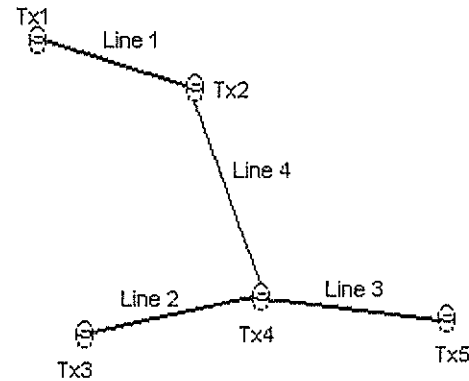


Figure 1 Figure showing why it is necessary to test for connectedness of a network.

In Fig 1 above, Lines 1, 2 and 3 were drawn using the straight-line distance list, as described in Step 2. However, that would result in two unconnected groups of transformers, with transformers 1 and 2 in a group and transformers 3, 4 and 5 in another group. A connectedness test would show that the two groups were not connected.

Step 4: The straight-line distance list must be examined to find the shortest line that will connect the two groups of transformers, as indicated by Line 4 in Fig. 1 above.

4. Transferring the straight-line network onto a CAD map

The straight-line network must now be transferred to a CAD map, and the straight lines must be rerouted so that they follow permissible cable routes. Design decisions must be made about where cables will be allowed to run. Typically, MV cables are routed along township roads and along stand boundaries, but they are not allowed to cross stands. This is the design strategy that was adopted for this software tool.

In previous research done by the Wits SED group, algorithms were developed for creating intelligent maps. Using these methods, an intelligent map would be created to 'recognise' a street block. A street block is defined as a boundary encompassing a group of

For the purposes of the automatic routing tool, the street blocks are seen as obstacles, which may not be crossed by MV cables.

An automated routing tool would have to be divert the MV cables around the obstacles.

From transformer Tx1, the straight line to transformer Tx2 will be followed, until an intersection with an obstacle is detected. The 8 vertices closest to the intersection point (vertices a, b, c, d, i, j, k and l) will then be found and the straight-line distance from each of those vertices to Tx2 will be calculated. The cable will then be routed to the vertex that is the shortest straight-line distance from Tx2, in this case vertex c. A check will be performed to ensure that the new line does not intersect the obstacle. This check would prevent the line from being routed to vertex d, for example.

This sequence of actions is repeated until Tx2 is reached. In the above drawing, the cable route would be Tx1→c→d→g→h→Tx2.

5. Conclusion

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Methods for calculating transient stability limit

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Abstract— This paper presents an overview of presently available methods for transient stability limit (TSL) calculation. In this overview we explain the classical methods based on time-domain simulation, which are in use today, as well as the new developments. It is apparent, that the classical time domain simulation can't meet the new requirements of the market-driven electricity system operation. Suitable for this new tasks are direct or hybrid methods, like transient energy function, equal area criterion and signal energy analysis. Furthermore the paper gives a comparison between the classical and the new approaches and between the several new methods.

I. INTRODUCTION

In many countries of the world the electricity market is undergoing a deregulation. This change will force the companies to work as efficient as possible. The companies will have to optimize the output of their power plants. Instead of building new power plants near the load centres, the energy will be transported from power plants with superfluous output to the load centres. This and the independent power producer will make the network more flexible and the changes will be less foreseeable for the system operator [1][2].

On the other hand, there is the trend to use the power system up to its maximal capacity, which leads to a more stressed network. To ensure the stability of this network, it is necessary to have calculation methods, that give accurate knowledge of the stability limits. This knowledge allows to increase the output of the network and leads to economical benefits.

To meet the new requirements of the stability assessment, new methods must be applied. Table 1 shows the requirements for the present and the future stability calculation tools. The greatest challenge is, that the stability is not done in advance anymore, but on-line. To meet the requirements of the real-time stability assessment direct and hybrid methods look most promising. These methods have up to now only been used for fast contingency screening. The present method to assess the stability limit is the time domain simulation. This method has superb modeling capa-

bilities and is furthermore used for system planning.

Table 1
Requirements for TSL calculation tools

	Present	Future
	Operation, Planning	Real-time
	Time-domain simulation	direct & hybrid methods
duration of study	not critical	critical
parameters and conditions	forecasted	measured
number of cases	large	one
number of scenarios	large	few
accuracy	not 100 %	100%
safety margin	large	small

The first attempts to speed up the stability limit calculation was to incorporate energy analysis into the time domain simulation to get a stability index [3]. Based on the Lyapunov's method the transient energy function (TEF) was developed at the Iowa State University, USA [4] [5]. Using this approach B. C. Hydro, Canada developed the Second-Kick-Methods [6], combining TEF and time domain simulation. This combination is as well used for the Hybrid and the Relevant-fault-on-trajectory (RFT) method, which was developed for the network of Ontario Hydro, Canada [2] [7]. The Ecole Polytechnique de Montreal concentrated on the development of the Signal Energy Analysis based on the Fourier methods [8][9]. The Powertech Labs Inc., Canada integrated the extended equal area criterion (EEAC) in an on-line dynamic security assessment [1].

This paper is structured as follows: Chapter II reviews the classification of the stability problem. Chapter III shows the presently used method to determine the TSL. Chapter IV concentrates on the new methods, which are supposed to meet the requirements of the new market. Chapter V concludes this paper.

II. STABILITY CLASSIFICATION

There are several classes in which stability problems can be subdivided. The stability classification shown in Figure 1 is built upon modeling purposes and it is concentrating on the large-signal rotor angle stability.

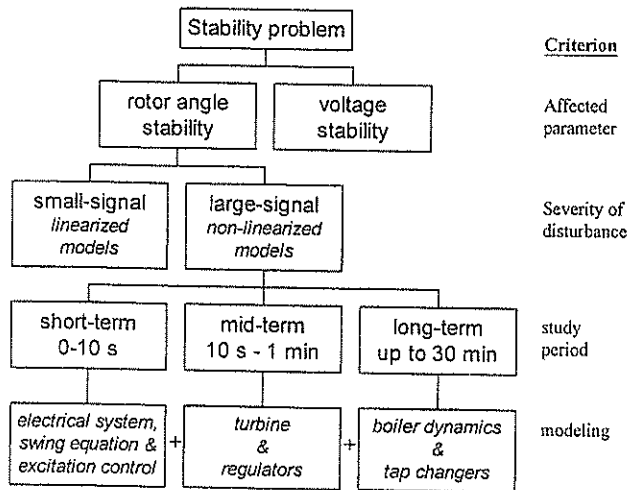


Fig. 1. Classification of the rotor angle stability problem

The first class is the subdivision in the affected parameter of the network. A progressive and uncontrollable drop of voltage after the occurrence of a contingency is called voltage instability. This instability occurs, when the system can't meet the demand of reactive power [10].

Monitoring the rotor angle of one or of a group of generators, the system is stable if relative rotor angles stay in certain limits. When a fault occurs on one of the connecting lines the generator can't push all of the energy it generates to the load. The generator has to store this superfluous energy in its moving elements. This causes the rotor to run faster and with that the rotor angle increases. If the fault is cleared before a critical time is reached the generator will come back to its prefault operation point. If the fault stays longer, than the generator will lose synchronism and it has to be disconnected from the system.

The small-signal stability is the ability of the power system to maintain synchronism under small disturbances. This disturbances lead to small perturbations around the equilibrium point. The reaction of the system can be calculated by using linearized models [10].

The large-signal stability is divided in short-term, mid-term and long-term stability. The system response, after a large disturbance, involves large excursions of generator rotor angles and it is influenced by the nonlinear power-angle relationship. Because

of that, the non-linear models must be used, and the time domain simulation or the direct methods incorporating non-linear models are only suitable.

The stability of interest in this report is the short-term stability, which study time is up to 10 s. It is, as well called transient stability. To simulate the response of the system after a disturbance for this short time, one only has to model the electrical system, swing equation and the excitation control. With a rising study period, there has to be incorporated the models of turbine, regulators, boiler dynamics, tap changers etc.

III. PRESENT UTILITY PRACTICE IN TSL CALCULATION

The dynamic security assessment is carried out off-line and it is done by the time domain simulation, which has been extensively developed over the last 20 years. Because of that it provides time responses of all quantities and has superb power system modeling capabilities. The simulation is run for a large number of forecasted system conditions and a specified set of contingencies. The answer for every simulation is a yes-or-no answer for the stability. The stability limit is derived out of a trial-and-error approach. Running the simulation for all the conditions and contingencies the stability limit derivation takes a long time and is usually carried out overnight. The results of the time-domain simulation are highly accurate, but paying attention to the fact, that all input parameters are forecasted and will not appear exactly like that, the safety margin must be large [2][11][13].

IV. EMERGING TOOLS FOR TSL CALCULATION TO BE USED IN MARKET-DRIVEN UTILITY OPERATION

To secure safe supply and the economic benefits, the new methods for TSL calculation have to be flexible, fast, accurate and they have to give information about the transient energy margin. The methods to meet these requirements are the transient energy function (TEF), the Equal Area Criterion (EAC) and the Signal Energy Analysis.

A. Transient Energy Function

The TEF is an extension of the Lyapunov's method. Lyapunov proposed, that as long as the energy function $V(x)$, which is the sum of the kinetic and potential energy of the postfault system is smaller than a critical value V_{cr} the system will be stable. The energy function used for TEF is the first integral of

motion of the system referred to a center of inertia. As long as you have a single-machine infinite-bus system, there is only one V_{cr} . For a multimachine system, however, one can find a stable equilibrium point (SEP) surrounded by a number of unstable equilibrium points (UEP), as shown in Figure 2. The arrow marks the potential energy boundary surface (PEBS), on which all the UEPs with its V_{cr} are lying. When a disturbance occurs, the system will leave the SEP and going toward the PEBS. As long as it doesn't cross the PEBS the system is stable.

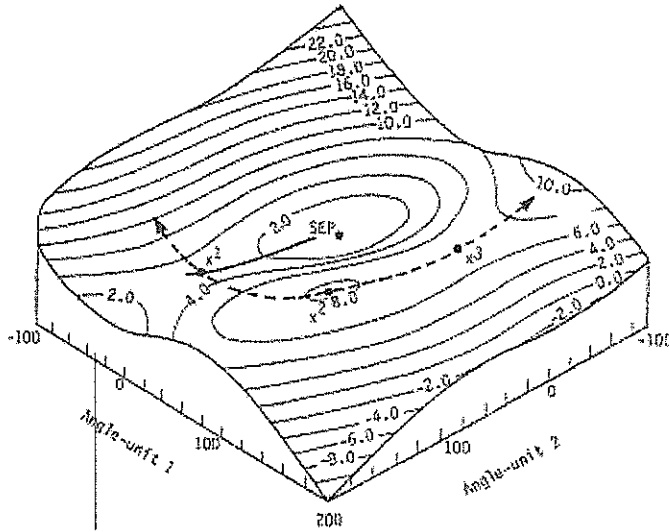


Fig. 2. PEBS for two machine system

The following equations show the energy function for a single-machine infinite-bus system for the post-fault system:

$$V(\delta, \omega) = \frac{1}{2}M\omega^2 - P_m(\delta - \delta^s) - P_e^{pf}(\cos \delta - \cos \delta^s) \quad (1)$$

$$= V_{KE} + V_{PE}(\delta, \delta^s) \quad (2)$$

with δ power angle,
 ω rotor speed,
 M inertia,
 P_m mechanical power of the system,
 δ^s power angle of a stable equilibrium point,
 P_e^{pf} maximum power of the postfault system,
 V_{KE} kinetic energy of the system and
 V_{PE} potential energy of the system.

For the time t_{cl} the fault is cleared, the energy func-

tion is:

$$V^{cl}(\delta^{cl}, \omega^{cl}) = \frac{1}{2}M\omega_{cl}^2 - P_m(\delta^{cl} - \delta^s) - P_e^{pf}(\cos \delta^{cl} - \cos \delta^s) \quad (3)$$

$$= V_{KE}^{cl} + V_{PE}^{cl} \quad (4)$$

The critical clearing time is reached, when

$$V_{PE}(\delta^u) - V_{PE}^{cl}(\delta^{cl}) = V_{KE}^{cl} \quad (5)$$

where δ^u is the rotor angle of the UEP and $V_{PE}(\delta^u)$ the critical energy V_{cr} . The computation of V_{cr} in the multimachine system is the challenging task of the TEF method and there are three different ways to do it, the Mode-of-disturbance method (MOD), the Exit method and the Sustained-fault method [12].

The MOD [6] method formulates a computational intensive multidimensional optimization problem and solves it by directly computing the UEP. Using this method and results from the time domain simulation the Second-Kick-Method has been applied at B.C. Hydro. More research must be done, because the method shows convergence problems, when the system is stressed.

The Exit method [11] simulates the sustained fault, till the trajectory reaches the PEBS. The boundary controlling UEP (BCU) method is derived out of this method, but the trajectory sometimes converges to the wrong UEP.

To avoid the convergence problem the Sustained-fault method [2] computes not V_{cr} , but the critical clearing time t_{cr} . This method has only a short coming, when the fault clearing time is much shorter than the critical clearing time. Much research has been done at Ontario Hydro based on the sustained-fault method. The developed Hybrid method has the disadvantage that the transient energy margin is not decreasing monotonically as the system conditions are increasingly stressed. The RFT (relevant fault-on trajectory) method remedies this shortcoming by using a corrected kinetic energy.

B. Equal Area Criterion

The EAC basically uses the faulted and postfault power angle curves to calculate the stability. Figure 3 shows the power-angle relationships.

The area A1 represents the kinetic energy and the area A2 the potential energy. The equations (6) and (7) describe the areas A1 and A2 in Figure 1.

$$A1 = \int_{\delta_0}^{\delta_c} (P_m - P_e^f \sin \delta) d\delta \quad (6)$$

$$A_2 = \int_{\delta_c}^{\delta_u} (P_e^{pf} \sin \delta - P_m) d\delta \quad (7)$$

with δ_0 power angle in steady state,
 δ_c power angle after the fault is cleared,
 P_m mechanical power of the system,
 P_e^f maximum power of the faulted system,
 P_e^{pf} maximum power of the postfault system,
 δ_u maximum power angle of the postfault system.

As long as the kinetic energy is smaller than the potential energy the system is stable [10].

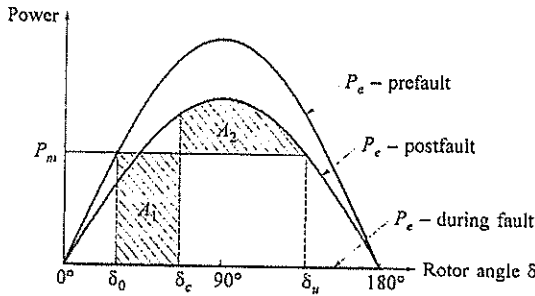


Fig. 3. Equal Area Criterion for SMIB system

The applied method based on EAC is called FASTEST (Fast Analysis of Stability using the Extended Equal Area Criterion and Simulation Technologies)[1]. It uses the time domain simulation to create the needed power angle curves. This method gives quantitative informations and additional to this it identifies the critical cluster of generators .

C. Signal Energy Analysis

The signal energy analysis [8][9] is built on the hypothesis, that the network impulse response h , monitored at the location j , as a function of time t and generation P carries information about the relative severity of the contingency, occurring at location i . The following integral (8) defines the signal energy:

$$E_{ij}(P) = \int_{-\infty}^{\infty} h_{ij}(t, P)^2 dt \quad (8)$$

When the network approaches instability the signal energy smoothly is running towards infinity. This defines the network stability limit for a particular contingency and makes the calculation of stability margins possible. As P is reaching the stability limit, the equation (8) can be approximated by:

$$E_{ij}(P) = \frac{C_{0ij}}{L_i - P} \quad (9)$$

Only two unknowns are in this relation, the coefficient C_{0ij} and the transfer limit L_i . These can be estimated by using results of the time domain simulation.

V. CONCLUSION

In this paper we identify the requirements for TSL calculation tools to be used in open-market utility operation. A number of recently developed tools are critically evaluated, against these requirements. Selection of the best method for the South African grid operation is the matter of the ongoing research.

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Iterative solution of large and sparse systems of linear equations appearing in the power system state estimation problem

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Abstract—Various iterative methods of solving a large system of linear equations will be discussed. The performance of these methods in solving large power system state estimation problems will be compared.

I. INTRODUCTION

Large sparse systems of linear equations of the form

$$\mathbf{Ax} = \mathbf{b} \quad (1)$$

frequently arise in power system state estimation problems. The normal equations, from the weighted least squares algorithm, of the form

$$(\mathbf{H}^T \mathbf{W} \mathbf{H}) \mathbf{x} = \mathbf{H}^T \mathbf{W} \mathbf{z} \quad (2)$$

are in form (1), where $\mathbf{H}$ is the Jacobian matrix and $\mathbf{W}$ is the matrix consisting of the weights assigned to each measurement taken from the power system. Methods to solve these systems of equations are classified into direct, stationary iterative and semi-iterative (or non-stationary iterative).

Direct solvers use variations of Gaussian elimination (e.g. LU factorization, where $\mathbf{A} = \mathbf{LU}$). A Newton-Raphson method employing LU factorization has been widely used in state estimation solution algorithms. However the LU factorization of the load flow Jacobian in each Newton iteration is a computationally expensive task. The LU factorization can be avoided by using an iterative method to solve the system of equations [1][2][3]. For large-scale power systems the LU factorization of the Jacobian consumes about 85% of the total Newton-Raphson process execution time.

Various iterative methods will be discussed in section II. For the sake of perspective and completeness, stationary iterative methods are briefly mentioned, although the subsequent investigation only considers non-stationary iterative methods. Methods to precondition a matrix, to get better numerical properties will be discussed in section III, and the sparsity of the systems of linear equations will be covered in section IV. In section V numerical results will be presented, from which a choice of best method is made.

II. ITERATIVE METHODS

A. Stationary iterative methods

The term *iterative method* refers to anyone of many techniques that use successive approximation to obtain the solution of a linear system of equations, given in *fixed-point*

form

$$\mathbf{x} = \mathbf{T}(\mathbf{x}) \quad (3)$$

Equation (3) can be solved choosing an initial approximation $\mathbf{x}_0$ and then calculating more accurate approximations by using the iteration

$$\mathbf{x}_{k+1} = \mathbf{T}(\mathbf{x}_k), \quad k = 1, 2, 3, \dots \quad (4)$$

An iteration (4) is called stationary if the definition of the operator $\mathbf{T}$ does not depend on the iteration step.

Let us split the matrix $\mathbf{A}$ into two matrices $\mathbf{B}$ and $\mathbf{C}$, so that $\mathbf{A} = \mathbf{B} - \mathbf{C}$. From (1) it then follows that

$$(\mathbf{B} - \mathbf{C})\mathbf{x} = \mathbf{b} \quad (5)$$

$$\mathbf{B}\mathbf{x} = \mathbf{C}\mathbf{x} + \mathbf{b} \quad (6)$$

The iterative method is

$$\mathbf{B}\mathbf{x}_{k+1} = \mathbf{C}\mathbf{x}_k + \mathbf{b} \quad (7)$$

where k is the index of iteration. The iterative step will be

$$\mathbf{x}_{k+1} = \mathbf{B}^{-1}\mathbf{C}\mathbf{x}_k + \mathbf{B}^{-1}\mathbf{b} \quad (8)$$

Let $\hat{\mathbf{x}}$ be the exact solution, $\mathbf{x}_k$ the estimate of $\hat{\mathbf{x}}$ at the k^{th} iteration and $\mathbf{x}_0$ the initial guess. The error at the k^{th} iteration is

$$\epsilon_k = \|\hat{\mathbf{x}} - \mathbf{x}_k\|, \quad (9)$$

where $\|\cdot\|$ indicates the norm of a vector. Equations (6) and (7) give the following relationship between ϵ_{k+1} and ϵ_k

$$\mathbf{B}\epsilon_{k+1} = \mathbf{C}\epsilon_k \quad (10)$$

This leads to

$$\epsilon_{k+1} = \mathbf{B}^{-1}\mathbf{C}\epsilon_k \quad (11)$$

and then

$$\epsilon_{k+1} = (\mathbf{B}^{-1}\mathbf{C})^{k+1}\epsilon_0 \quad (12)$$

where $\epsilon_0 = \|\hat{\mathbf{x}} - \mathbf{x}_0\|$. The iterates of $\mathbf{x}_k$ will converge to the true solution if the spectral radius of $\mathbf{B}^{-1}\mathbf{C}$ is smaller than one. The spectral radius of a matrix is defined to be the largest eigenvalue of that matrix.

Let $\mathbf{A} = \mathbf{D} - \mathbf{L} - \mathbf{U}$, where $\mathbf{D}$ is the diagonal of $\mathbf{A}$, $\mathbf{L}$ is a strictly lower triangular matrix whose elements are the negative of elements of $\mathbf{A}$ that are below the diagonal and

U is a strictly upper triangular matrix whose elements are the negative of elements of A that are above the diagonal.

The following methods belong to the class of stationary iterative methods:

• **Jacobi Method:**

$$B = D \text{ and } C = L + U, \text{ giving}$$

$$Dx_{k+1} = (L + U)x_k + b \quad (13)$$

This method converges very slowly.

• **Gauss-Seidel method:**

$$B = D - L \text{ and } C = U, \text{ giving}$$

$$(D - L)x_{k+1} = Ux_k + b \quad (14)$$

This method converges faster than the Jacobi method, but is still inefficient.

• **Successive Over-Relaxation (SOR) method:**

$$B = D - \omega L \text{ and } C = (1 - \omega)D + \omega U, \text{ giving}$$

$$(D - \omega L)x_{k+1} = [(1 - \omega)D + \omega U]x_k + \omega b \quad (15)$$

If ω (relaxation parameter) is chosen well, then this method will have a good convergence rate [4]. The choice of the optimal ω depends on the spectral radius of the matrix $B_J = D^{-1}(L + U)$, used in the Jacobi method.

$$\omega_{opt} = \frac{2}{1 + \sqrt{1 - \rho(B_J)^2}} \quad (16)$$

$$\rho(B_J) = \text{maximum eigenvalue of } [D^{-1}(L + U)] \quad (17)$$

where $\rho(B_J)$ is the spectral radius of B_J .

B. Non-stationary iterative methods

An iteration $x_{k+1} = T_k(x_k)$ is called non-stationary if the specification T_k depends on the iteration step. There are various non-stationary iterative methods for symmetric and non-symmetric matrices.

• **Conjugate Gradient (CG) Method:** This method is efficient for solving symmetric positive definite linear systems. Many of the other methods depend on the CG method. The iterates x_k are updated at each step by a multiple α_k of the search direction vector p_k :

$$x_k = x_{k-1} + \alpha_k p_k \quad (18)$$

The residuals $r_k = Ax_k - b$ are updated according to

$$r_k = r_{k-1} + \alpha_k q_k \quad \text{where} \quad q_k = Ap_k$$

and

$$\begin{aligned} \alpha_k &= \frac{r_{k-1}^T r_{k-1}}{p_k^T A p_k} \\ p_k &= -r_{k-1} - \beta_{k-1} p_{k-1} \\ \beta_{k-1} &= \frac{r_{k-1}^T r_{k-1}}{r_{k-2}^T r_{k-2}} \end{aligned}$$

This method takes about n steps to obtain a solution, where n is the size of the system.

• **Conjugate Gradient Method on the Normal equations (CGNR):** By applying the CG method to the equation $A^T A x = A^T b$, that is replace A with $A^T A$ and b with $A^T b$, one gets the CGNR method. The convergence rate may be unacceptable if the condition number of A , $[\kappa(A)]$ is not very close to *one*. The condition number is defined as

$$\kappa(A) = \left| \frac{\lambda_{\max}}{\lambda_{\min}} \right| \quad (19)$$

• **Generalized Minimal Residual (GMRES) method:** Form an orthogonal basis for the set of vectors, $\{r_0, Ar_0, A^2 r_0, \dots\}$, that defines the Krylov space, by using the next algorithm:

$$w_k = A v_k$$

$$\text{do } i = 1, 2, \dots, k$$

$$w_k = w_k - \langle w_k, v_i \rangle v_i$$

end do

$$v_{k+1} = \frac{w_k}{\|w_k\|}$$

where $\langle \cdot, \cdot \rangle$ defines the inner product of two vectors, $r_0 = b - Ax_0$, k is the iteration index and $v_1 = \frac{r_0}{\|r_0\|_2}$. The iterates are constructed as

$$x_k = x_0 + y_1 v_1 + y_2 v_2 + \dots + y_k v_k$$

where the coefficients y_k have been chosen to minimize $\|b - Ax_k\|$.

This method takes about n steps to obtain a solution. A drawback of this method is that the amount of storage required per iteration rises linearly with the iteration count.

• **Bi Conjugate (BiCG) Method:** It is the same as the CG method, except that two types of residuals and searching directions are used:

$$r_k = r_{k-1} + \alpha_k A p_k, \quad \tilde{r}_k = \tilde{r}_{k-1} + \alpha_k A^T \tilde{p}_k$$

$$p_k = -r_{k-1} - \beta_{k-1} p_{k-1}, \quad \tilde{p}_k = -\tilde{r}_{k-1} - \beta_{k-1} \tilde{p}_{k-1}$$

$$\alpha_k = \frac{\tilde{r}_{k-1}^T r_{k-1}}{\tilde{p}_k^T A p_k}, \quad \beta_{k-1} = \frac{\tilde{r}_{k-1}^T r_{k-1}}{\tilde{r}_{k-2}^T r_{k-2}}$$

Convergence may often be irregular, and sometimes the method may diverge from the solution.

III. PRECONDITIONING

The choice of a preconditioner involves the selection of a matrix Q , such that the preconditioned system is better conditioned than the original system, $Ax = b$. Preconditioning methods are:

• **Left preconditioning**

$$Q^{-1}Ax = Q^{-1}b \quad (20)$$

- Right preconditioning

$$(AQ^{-1})(Qx) = b \quad (21)$$

- Two-sided preconditioning

$$(Q_L^{-1}AQ_R^{-1})(Q_Rx) = Q_L^{-1}b \quad (22)$$

where Q is factorized as $Q = Q_L Q_R$.

One requirement for Q is that it must be easily invertible [5]. The preconditioned system has the same solution as the original system, but has a better condition number. Q must be chosen close to A , and still be easily invertible.

An iterative method can be preconditioned according to the following scheme [4]:

1. The transformation of the right-hand side: $b \rightarrow M^{-1}b$,
2. The application of the unpreconditioned iterative method to the linear system with the modified coefficient matrix $Q_L^{-1}AQ_R^{-1}$,
3. The computation of $x = Q_R^{-1}y$, where y is the solution resulting from the previous step.

There are various types of preconditioners:

- **Point Jacobi preconditioner:** $Q = D$, the diagonal of A .
- **Neumann Polynomial method:** A truncated Neumann series approximation to A^{-1} is used. The degree of the polynomial to be used must also be decided. Split A into $A = D - C$, where D is the diagonal of A , then $A = (I - CD^{-1})D$ so that

$$\begin{aligned} A^{-1} &= D^{-1}(I - CD^{-1})^{-1} \\ &= D^{-1}[I + CD^{-1} + (CD^{-1})^2 + \dots], \end{aligned}$$

where I is the unity matrix. A truncated form of this series is then used for Q^{-1} [5].

- **Incomplete Cholesky preconditioner:** The idea behind this approach is to find a lower triangular matrix H that is close to A 's exact Cholesky factor G , where $A = GG^T$, and G is lower triangular. A must be symmetric to have a Cholesky factor. To calculate H , do the normal Cholesky factorization, but let $h_{ij} = 0$ if the corresponding $a_{ij} = 0$. Use the following algorithm [7]:

```

for k = 1, 2, ..., n
  A(k, k) = sqrt(A(k, k))
  for i = k + 1, 2, ..., n
    if A(i, k) ≠ 0
      A(i, k) = A(i, k) / A(k, k)
    end
  end
  for j = k + 1, 2, ..., n
    for i = 1, 2, ..., n
      if A(i, j) ≠ 0
        A(i, j) = A(i, j) - A(i, k) A(j, k)
      end
    end
  end
end
end

```

- **Successive Overrelaxation preconditioner: (SOR):** Split A into $A = D - C_L - C_U$, where D is the diagonal of A , C_L is a lower triangular matrix, C_U is an upper triangular matrix and ω is the same as for the SOR iterative scheme. Q is then defined as [5]:

$$Q = \frac{1}{\omega}D - C_L$$

IV. SPARSITY

The matrix H^TWH from equation (2) is sparse. This is due to the fact that not all the nodes in the power system are connected to each other.

There are several schemes available to store these sparse matrices. The computer package, NSPCG (Nonsymmetric Preconditioned Conjugate Gradient) used to simulate the numerical results, supports some of these storage schemes. The storage method used during the simulation is called the *primary storage*. Consider the following example:

In this storage format, there are two rectangular arrays COEF and JCOEF, one real and one integer, which are used to store a representation of the coefficient matrix. Both arrays are dimensioned NDIM by MDIM, where NDIM is at least as large as N , the size of the system, and MDIM is at least as large as MAXNZ, the maximum number of nonzeros per row in the matrix, with the maximum being taken over all rows. Each row in COEF will contain the nonzeros of the corresponding row in the full matrix A , and the corresponding row in JCOEF will contain its respective column numbers.

For example, the matrix

$$A = \begin{pmatrix} 11 & 0 & 0 & 14 & 15 \\ 0 & 22 & 0 & 0 & 0 \\ 0 & 0 & 33 & 0 & 0 \\ 14 & 0 & 0 & 44 & 45 \\ 15 & 0 & 0 & 45 & 55 \end{pmatrix}$$

would be represented in the COEF and JCOEF arrays as

$$\text{COEF} = \begin{pmatrix} 11 & 14 & 15 \\ 22 & 0 & 0 \\ 33 & 0 & 0 \\ 44 & 14 & 45 \\ 55 & 15 & 45 \end{pmatrix}$$

$$\text{JCOEF} = \begin{pmatrix} 1 & 4 & 5 \\ 2 & 0 & 0 \\ 3 & 0 & 0 \\ 4 & 1 & 5 \\ 5 & 1 & 4 \end{pmatrix}$$

If a row in matrix A has fewer than MAXNZ nonzeros, the corresponding rows in COEF and JCOEF should be padded with zeros [5].

In direct methods the matrix elements must be re-ordered, during the process of Gaussian elimination and LU factorization, to preserve the sparsity of the matrix [8]. In iterative methods reordering is not necessary and this is an advantage of iterative methods over direct methods.

V. NUMERICAL RESULTS

Systems of equations, that are close to the equations in practice, were simulated and the time of each method to converge to the solution, were taken. The various preconditioners were also tested with the methods.

Jacobi Preconditioner	CGNR	GMRES	BiCG
Number of equations	Time (s) to converge		
25	0.33	0.27	0.22
100	0.38	0.33	0.33
225	0.49	0.38	0.39
400	0.60	0.55	0.55
625	0.82	0.71	0.66
900	0.99	0.94	0.93
1225	6.31	6.93	5.55
1600	8.52	14.00	9.50
2025	11.09	17.96	11.65
2500	15.49	19.23	22.08

Table 1: Jacobi preconditioner

In table 1 the results are shown for the Jacobi preconditioner applied to the CGNR, GMRES and BiCG semi iterative methods. The number of equations are shown in the left column and the other three columns show the time (in seconds) taken by each method to converge to the solution. Only left preconditioning was implemented, because some of the methods in the NSPCG package do not support right or two-sided preconditioning. The stopping criteria used to test for convergence was

$$\left[\frac{\langle z_n, z_n \rangle}{\langle Q_L^{-1} b, Q_L^{-1} b \rangle} \right]^{\frac{1}{2}} < \zeta, \quad (23)$$

where $\langle \cdot, \cdot \rangle$ represents the inner product between two vectors, $z_n = Q_L^{-1} r_n$ is the left-preconditional residual, $r_n = b - Ax_n$ is the current residual and x_n is the current iterate. ζ was chosen to be 10^{-10} . Table 2 and 3 are the same as table 1, but they show the incomplete Cholesky and Neumann polynomial preconditioners respectively applied to the CGNR, GMRES and BiCG semi iterative methods. The order of the Neumann polynomial used, is one. In the NSPCG package the SOR preconditioner may only be applied to the SOR iterative method, and the SOR preconditioner is therefore not used in the simulations.

Cholesky Preconditioner	CGNR	GMRES	BiCG
Number of equations	Time (s) to converge		
25	0.33	0.27	0.27
100	0.9	0.33	0.33
225	0.50	0.44	0.44
400	0.71	0.60	0.60
625	0.93	0.82	0.76
900	1.21	1.05	1.00
1225	8.63	8.68	7.75
1600	13.40	13.29	12.58
2025	14.72	8.84	9.01
2500	16.37	17.41	32.52

Table 2: Incomplete Cholesky preconditioner

Neumann Preconditioner	CGNR	GMRES	BiCG
Number of equations	Time (s) to converge		
25	0.33	0.28	0.27
100	0.39	0.33	0.33
225	0.55	0.44	0.44
400	0.77	0.66	0.66
625	0.99	0.88	0.82
900	1.32	1.15	1.10
1225	6.54	3.79	4.23
1600	8.35	8.02	9.61
2025	10.60	10.00	10.66
2500	14.01	12.25	19.72

Table 3: Neumann polynomial preconditioner

From table 3 it can be seen that the best performance for solving the systems of equations was obtained by applying the Neumann polynomial preconditioner to the GMRES method. For comparison of direct methods with iterative methods, see [3]. As the system size increases, the iterative methods give better performance than the direct methods.

VI. CONCLUSIONS

- This paper presented a discussion on different stationary and non-stationary iterative methods. The matrix $H^T W H$ of equation (2) is symmetric, but there are other formulations of the power system state estimation problem that do not produce symmetric matrices, and then iterative methods and preconditioners that do not require symmetric matrices can be used to solve symmetric and non-symmetric problems.
- Direct methods, like LU factorization, are very inefficient compared to iterative methods, for very large power systems.
- The iterative method that gave the best performance in solving the systems of equations was the GMRES method with the Neumann polynomial preconditioner applied to it.

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Application of Total Least Squares in Power System State Estimation

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Abstract— In this paper we investigate the application of the Total Least Squares method as an optimal state estimation method for time varying model parameters. Comparison between Least Squares and Total Least Squares is presented in the paper.

gives rise to an *overdetermined* set of m linear equations ($m > n$) :

$$\mathbf{Ax} = \mathbf{b} \quad (1)$$

or

$$\begin{aligned} a_{11}x_1 + a_{12}x_2 + \dots + a_{1n}x_n &= b_1 \\ a_{21}x_1 + a_{22}x_2 + \dots + a_{2n}x_n &= b_2 \\ &\vdots \\ a_{m1}x_1 + a_{m2}x_2 + \dots + a_{mn}x_n &= b_m \end{aligned}$$

I. INTRODUCTION

The Total Least Squares (TLS) method has been introduced in recent years in the numerical analysis literature as an alternative for Least Squares (LS) method in the case where all data is effected by errors. This paper reviews the main principle of the TLS method in solving overdetermined sets of linear equations $\mathbf{Ax} \approx \mathbf{b}$ appearing in Power System State Estimation.

State Estimation is the process of assigning a value to an unknown system state variable based on measurements from that system according to some criteria. Usually, the process involves imperfect measurements that are redundant and the process of estimating the system states is based on a statistical criterion that estimates the true value of the state variables to minimize or maximize the selected criterion. A commonly used and familiar criterion is that of minimizing the sum of the squares of the differences between the estimated and "true" (i.e. measured) values of a function.

In a power system, the state variables are the voltage magnitudes (V) and relative phase angles (δ) at the system nodes. Measurements are required in order to estimate the system performance in real time for both system security control and constraints on economic dispatch. The inputs to an estimator are imperfect power system measurements of voltage magnitudes and power, VAR, or ampere-flow quantities. Once the system voltage and phase angles of a system are estimated, and knowing the line impedances, the line flows and bus injections can be calculated [1].

In section II, the problem formulation for the TLS State Estimation problem will be pointed out. To solve for the state variables, the non-linear power system equations must first be linearized in order to obtain a system of equations of the form $\mathbf{Ax} \approx \mathbf{b}$. The SVD (singular value decomposition) will be used as the solution method which will be discussed in section II-B. Numerical results will be shown in section III and a conclusion will be given at the end.

II. TOTAL LEAST SQUARES STATE ESTIMATION

A. Total Least Squares method

A basic problem of TLS is to determine an estimate of the *unknown* parameters from certain measurements. This

where $\mathbf{A} \in R^{m \times n}$ is the data matrix and $\mathbf{b} \in R^m$ is the observation vector.

In the least squares (LS) approach the measurements of the data matrix $\mathbf{A}$ are assumed to be free of error. All errors are confined to the observation vector $\mathbf{b}$ in the right-hand side of (1) [2].

The assumption for no errors of the data matrix $\mathbf{A}$ is not always true. Total least squares (TLS) is one method that is appropriate when there are errors in *both* the observation vector $\mathbf{b}$ and the data matrix $\mathbf{A}$ of (1). An application where TLS would be appropriate in Power System State Estimation, is for example for transmission lines where the conductor resistance varies with temperature. This can be modeled by the following formula

$$R_2 = R_1 [1 + \alpha(t_2 - t_1)] \quad (2)$$

R_2 is the resistance at temperature t_2 , R_1 is the resistance at temperature t_1 and α is the temperature coefficient. Thus a change in line resistance will have an effect on data matrix $\mathbf{A}$.

The TLS problem can be outlined as follows: Given an *overdetermined* set of m linear equations ($m > n$), the TLS problem minimize the correction applied to the data matrix $\mathbf{A}$ and the observation vector $\mathbf{b}$ by using the Frobenius norm. The Frobenius norm of a $m \times n$ matrix $\mathbf{A}$ is defined by :

$$\|\mathbf{A}\|_F = \sqrt{\sum_{i=1}^m \sum_{j=1}^n a_{ij}^2} \quad (3)$$

Once such a solution is found which satisfy (1), the solution is called a TLS solution [3].

B. Problem Formulation

The following types of measurements occur in a power system network :

- Flows - Real and reactive power flows measured at the terminal buses of a transmission line or transformer.
- Injections - Real and reactive power net injections at

system buses.

- Voltage magnitude - At system buses.
- Current magnitude - Ampere flows along transmission lines or transformers measured at the terminal buses.

If there are m measurements, then these measurements can be written as a m -dimensional vector Z_M :

$$Z_M = Z_T + v \quad (4)$$

where

- Z_M - measurement as received from a measurement device (voltage, current, active and reactive power, phase angle)
- Z_T - true value of the quantity being measured.
- v - random measurement error (high frequency noise, drift, calibration errors, analog to digital errors etc.)

In a power system, the non-linear equations for the power flows and power injections must first be linearized by using the Taylor series in order to solve for the state variables by using TLS. This could be done using the observation matrix equation

$$Z_M = h(X) + v \quad (5)$$

where $h(X)$ is the nonlinear vector function relating the measurement vector to the state vector and v is a vector containing the errors. $h(X)$ in (5) must be linearized by using the Taylor series expansion :

$$Z_M = h(X_k) + \left. \frac{\partial h(X)}{\partial X} \right|_{X_k} (X_{k+1} - X_k) + v \quad (6)$$

$$Z_M - h(X_k) = H(X_k)(X_{k+1} - X_k) \quad (7)$$

$$\Delta Z_M = H \Delta X \quad (8)$$

where

$\Delta Z_M = Z_M - h(X_k)$: The difference between the measurement vector and the power flows and injection equations for a power system network.

$H = H(X_k)$: The Jacobian matrix.

$\Delta X = X_{k+1} - X_k$: The state correction vector containing the state variables (V, δ) .

k - iteration counter.

We obtain a system of equations of the form $Ax \approx b$ as can be seen from equation (8) where ΔZ_M is equivalent to the observation vector b and matrix H is equivalent to the data matrix A . We can now apply the TLS problem to (8) and solve for the state correction vector ΔX iteratively, by using the singular value decomposition (SVD).

C. Solution Method

The Singular Value Decomposition (SVD) is of great practical importance for solving TLS problems. In practice, an equation of the form (8), is solved using SVD to avoid problems due to possible ill-conditioning of the matrix A or in our case, matrix H . The SVD of a $m \times n$ matrix A , $m > n$, gives the following decomposition:

$$A = U \times S \times V^T \quad (9)$$

Matrix S is a diagonal matrix of the same dimension as A and with non-negative diagonal elements in decreasing order. Matrices U and V are unitary matrices with dimensions $m \times m$ and $n \times n$ respectively.

For the TLS problem, the following algorithm can be used for solving a system of equations of the form $Ax \approx b$ by using the SVD:

Given $A \in R^{m \times n}$ and $b \in R^m$.

Step1 : Compute the SVD, i.e. $[A; b] = U \times S \times V^T$.

Step2 : If $v_{n+1, n+1} \neq 0$ of matrix V , then

$x = -1/(v_{n+1, n+1}) \times [v_{1, n+1}, \dots, v_{n, n+1}]^T$ is the TLS solution.

If $v_{n+1, n+1} = 0$ of matrix V and the previous diagonal element of matrix S , s_n , is greater than s_{n+1} , then TLS fails to have a solution [3].

As can be seen from *step 1*, the dimension of matrix S will then be the same as $[A; b]$ and the dimension of matrix V increase to $(n+1) \times (n+1)$ because of computing the SVD of $[A; b]$ and not just of matrix A . The next example will illustrate the algorithm for obtaining a TLS solution.

Let $m = 3$ and $n = 2$.

Let $A = \begin{bmatrix} 1 & 2 \\ 1 & 1 \\ -2 & -4 \end{bmatrix}$ and $b = \begin{bmatrix} 1 \\ -1 \\ -2 \end{bmatrix}$.

Step 1 : The SVD of $[A; b]$ will be $[U, S, V] = \text{svd}([A; b])$

where

$$U = \begin{bmatrix} 0.4415 & 0.0712 & 0.8944 \\ 0.1593 & -0.9872 & 0.0000 \\ -0.8830 & -0.1425 & 0.4472 \end{bmatrix}$$

$$S = \begin{bmatrix} 5.5427 & 0 & 0 \\ 0 & 1.5095 & 0 \\ 0 & 0 & 0 \end{bmatrix}$$

$$V = \begin{bmatrix} 0.4270 & -0.4181 & -0.8018 \\ 0.8253 & -0.1822 & 0.5345 \\ 0.3695 & 0.8899 & -0.2673 \end{bmatrix}$$

Step 2 : $v_{3,3} = -0.2673 \neq 0$

$$-1/v_{3,3} = -1/(-0.2673) = 3.7411$$

The TLS solution is then obtained as

$$x_1 = 3.7411 \times -0.8018 = -3.00$$

$$x_2 = 3.7411 \times 0.5345 = 2.00$$

III. NUMERICAL RESULTS

LS and TLS were applied to a 3-bus system network and a 14-bus system network. Figure 1 illustrates the 3-bus system network and the following measurements (in per unit) were obtained for the 3-bus network by load flow analysis :

Bus nr.	2	3
P injections	-1	0.6506
Q injections	-0.2	0.1586

Bus nr.	1 - 2	1 - 3	2 - 3
P flows	0.4609	-0.1109	-0.5393
Q flows	0.1708	0.0027	-0.0775

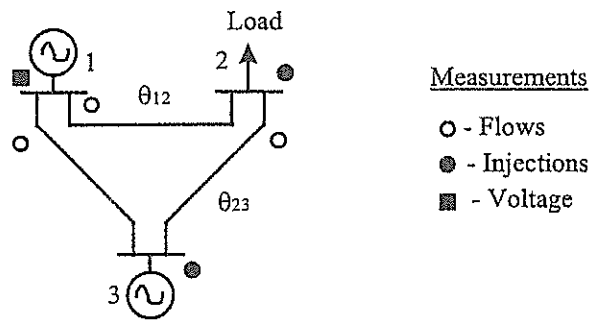


Fig. 1. 3-bus system network

Bus nr.	1	Bus nr.	1 - 2	2 - 3
Voltage	1	Teta (deg)	5.4445	-7.9875

The network data for the 3-bus system are the following (in per unit):

Bus nr.	1 - 2, 2 - 1	1 - 3, 3 - 1	2 - 3, 3 - 2
Resistance	0.2	0.4	0.25
Reactance	0.001	0.001	0.001

Bus nr.	1	2	3
Pgen	0.35	—	0.6506
Qgen	0.1735	—	0.1586
Pload	—	1	—
Qload	—	0.2	—

The figure and network data for the 14-bus network system can be found in [4].

The errors that are added at random to the data matrix A (in our case the parameter matrix H) and the observation vector b (in our case the vector ΔZ_M) are of a *normal* or *Gaussian* distribution with mean $\mu = 0$ and δ^2 the variance. Three cases are simulated for the 3-bus and 14-bus network.

- The first case is where the errors are of the same order for both the matrix A and the vector b .
- The second case is where the order of error is different. The order of error will be bigger for matrix A than that for vector b .
- The third case is where the order of error is also different but with bigger errors added to vector b than to matrix A .

The order of an error can be explained by the following example.

Let $a_{11} = 0.2$ and $b_1 = 0.6506$. Let the order of an error be maximum 15% of a_{11} and b_1 . A random number must

be generated with a value smaller or equal to 0.15. If the random number is for example 0.12, then the new values for a_{11} and b_1 with errors included will be :

$$a_{11\text{new}} = a_{11} + a_{11} \times 0.12 = 0.2 + 0.2 \times 0.12 = 0.224$$

$$b_{1\text{new}} = b_1 + b_1 \times 0.12 = 0.6506 + 0.6506 \times 0.12 = 0.7286$$

The criteria that will be used to compare the performance between LS and TLS, is to determine a *performance index* which is the norm of the difference between the state variables which are obtained from load flow, and the state variables containing the errors, which are obtained from LS and TLS.

In each of the three cases the simulation to obtain the state variables, will be repeated five hundred times for the 3-bus and 14-bus network. With each simulation the *performance index* is calculated. The number of times that LS or TLS has the smallest value for the *performance index* is then counted. The counted number is converted to a percentage value to indicate for each case whether LS or TLS is the best method to be used.

A. Case 1 : Errors are of the same order

3-bus system

The order of error that is added to matrix A and vector b is maximum 5% of the values of matrix A and vector b . The percentage value for LS having the smallest *performance index* was 42%. The percentage value for TLS having the smallest *performance index* was 58%.

The order of error that is added to matrix A and vector b is maximum 15% of the values of matrix A and vector b . The percentage value for LS having the smallest *performance index* was 34%. The percentage value for TLS having the smallest *performance index* was 66%.

14-bus system

The order of error that is added to matrix A and vector b is maximum 5% of the values of matrix A and vector b . The percentage value for LS having the smallest *performance index* as well as TLS having the smallest *performance index* were equal with each 50%.

The order of error that is added to matrix A and vector b is maximum 10% of the values of matrix A and vector b . The percentage value for LS having the smallest *performance index* was 39%. The percentage value for TLS having the smallest *performance index* was 61%.

B. Case 2 : Order of error is bigger for matrix A

3-bus system

The order of error is different. The order of error for matrix A is maximum 50% of the values of matrix A and the order of error for vector b is maximum 10% of the values of vector b . The percentage value for LS having the smallest *performance index* was 54%. The percentage value for TLS having the smallest *performance index* was 46%.

14-bus system

The order of error is also different. The order of error for matrix A is maximum 50% of the values of matrix A and the order of error for vector b is maximum 3% of the values of vector b . The percentage value for LS having

the smallest *performance index* was 60%. The percentage value for TLS having the smallest *performance index* was 40%.

C. Case 3 : Order of error is bigger for vector $\mathbf{b}$

3-bus system

The order of error is different. The order of error for vector $\mathbf{b}$ is maximum 60% of the values of vector $\mathbf{b}$ and the order of error for matrix $\mathbf{A}$ is maximum 10% of the values of matrix $\mathbf{A}$. The percentage value for LS having the smallest *performance index* was 46%. The percentage value for TLS having the smallest *performance index* was 54%.

14-bus system

The order of error is also different. The order of error for vector $\mathbf{b}$ is maximum 40% of the values of vector $\mathbf{b}$ and the order of error for matrix $\mathbf{A}$ is maximum 10% of the values of matrix $\mathbf{A}$. The percentage value for LS having the smallest *performance index* was 78%. The percentage value for TLS having the smallest *performance index* was 22%.

IV. CONCLUSIONS

From the numerical results, the following conclusions :

- If the errors are of the same order and very small, then TLS and LS give similar results. If the order of error is the same but bigger, then TLS will give better estimates than LS.
- In the case that the order of error differ in the measurement vector $\mathbf{b}$ and the parameter matrix $\mathbf{A}$, it seems that LS gives better estimates than TLS as for the 3-bus system. For the 14-bus system it is very clear that LS gives better estimates than TLS.
- To conclude, it is better to use TLS in cases where the errors are of the same order for vector $\mathbf{b}$ as well as matrix $\mathbf{A}$ and use LS in cases when there is a big difference in the order of error for vector $\mathbf{b}$ and matrix $\mathbf{A}$.

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APPLICATION OF QUANTITATIVE FEEDBACK THEORY TO SQUIRREL-CAGE INDUCTION MACHINE DRIVES

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Abstract This paper presents a design of feedback controllers using Quantitative Feedback Theory for a high performance drive system, to overcome the undesirable effects of uncertainty within a range of machines. A parameter estimator and a power efficiency search algorithm is also presented. It is demonstrated that by accurately quantifying the uncertainty within a plant specific dynamic performance at critical frequencies can be achieved. Digital simulation results are presented to substantiate the analysis.

INTRODUCTION

With processing power no longer a limitation modern variable speed drives are expected to satisfy stiff requirements such as fast speed and flux tracking and robustness against parameter uncertainties and load disturbance. Uncertainty in the response introduced when the plant used in the drive is different from the design model causes performance degradation. In this paper emphasis on making the drive universal for any machine within a specific power range with different plant parameters is demonstrated.

Uncertainty in the current measurements and flux estimates also has a direct influence on the dynamic performance of the machine. In an attempt to reduce the cost of expensive electronics and maintenance to flux sensors, drive technology has moved away from direct flux control to indirect flux control; and in so doing makes the performance more sensitive to changes in the machine parameters.

This paper outlines the effects of uncertainty on the performance of a squirrel-cage induction machine, and shows how Quantitative Feedback Theory, QFT can be applied to improve the dynamic performance of such systems.

An alternate decoupling technique suggested by Kim[1] was investigated and used as a basis to design the feedback controllers. Other work investigated regarding vector control were Liaw[2] and Brahim[3]. A five state induction machine is modeled in the frequency domain using five different machines chosen from the literature surveyed to quantify the uncertainty of the range of machines.

A rotor resistance estimator has also been developed on the premise that the stator power is directly influenced by changes in the rotor parameters. This also reduces the uncertainty in the flux estimates as the flux changes with changes in the rotor resistance and inductance. A flux estimator commonly found in the literature for indirect flux control has been included in the control system.

Although performance and stability have been the main design criteria for the control system, energy saving which is lacking in the literature surveyed, has been included in the system. Kim's work on this issue was investigated. His results were verified and the technique was incorporated in the control system.

OVERVIEW OF THE SYSTEM

The major components of a general speed control system is shown in fig. 1. The control system consists of the speed and square-flux (sum of the squares of the d- and q-axis fluxes) feedback controllers; decoupler; current controllers; parameter estimator and the power efficiency controller.

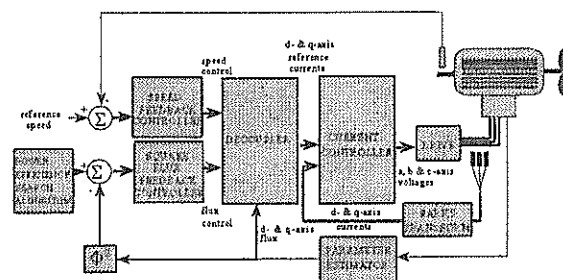


Fig.1 Squirrel-Cage Induction Machine Drive System

The speed and square-flux controllers ideally produce the optimum control signals for dynamic and steady state operation. The decoupler is on-line and attempts to decouple the speed and square-flux dynamically. Tight loops around the current controller ensures that the current error is kept as close to zero as possible. The plant used is the five state unmodified induction machine model in the stationary dq-axis reference frame; the external inputs are the reference speed; the measured speed and estimated square-flux are the control variables.

MODELING OF A SQUIRREL-CAGE INDUCTION MACHINE IN THE FREQUENCY DOMAIN

The plant was modeled in the frequency domain for the design on the Nichols chart. A general two channel, speed-flux controller is presented in fig. 2.

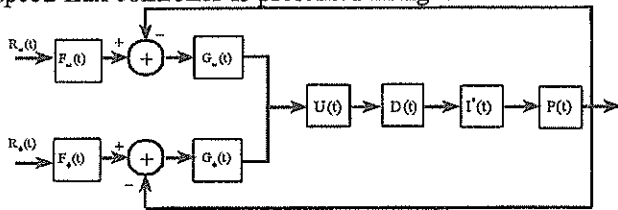


Fig. 2. Block Diagram of the Control System

The reference signal $R_\omega(t)$ is driven by the user and $R_\phi(t)$ by the power efficiency search algorithm. $F_\omega(t)$ and $F_\phi(t)$ are the prefilters to the system. The high order feedback controllers $G_\omega(t)$, and $G_\phi(t)$ produce $\underline{U}(t)$, the control vector carrying the decoupling complexities. These signals pass through the decoupler $D(t)$. The result is the reference dq-current vector and its difference from the actual stator current which drives a bang-bang hysteresis current controller, $I^*(t)$. $I^*(t)$ produces the reference voltage signals in the dq-axis frame to drive the machine, $P(t)$ to the requested speed and square-flux.

Kim's decoupling technique is dynamic and non-linear; it is different from Field Oriented Control in that the speed and square-flux are decoupled directly and not the d- and q-axis. The decoupler is presented in eq. 1.

$$\begin{bmatrix} g_\omega(t) \\ g_\phi(t) \end{bmatrix} = \begin{bmatrix} -\phi_q(t) & \phi_d(t) \\ \phi_d(t) & \phi_q(t) \end{bmatrix} \begin{bmatrix} i_d(t) \\ i_q(t) \end{bmatrix} \quad (1)$$

$g_\omega(t)$ and $g_\phi(t)$ are control signals from the feedback controllers and $\phi_d(t)$ and $\phi_q(t)$, and $i_d(t)$ and $i_q(t)$ are the d- and q-axis fluxes and currents respectively.

The combined complexities of the decoupler, current controller and plant were reduced to a 2×2 matrix structure and described only by a set of frequency response curves shown in fig. 3. Using Matlab with Simulink the system was simulated with no feedback to quantify each term in the matrix.

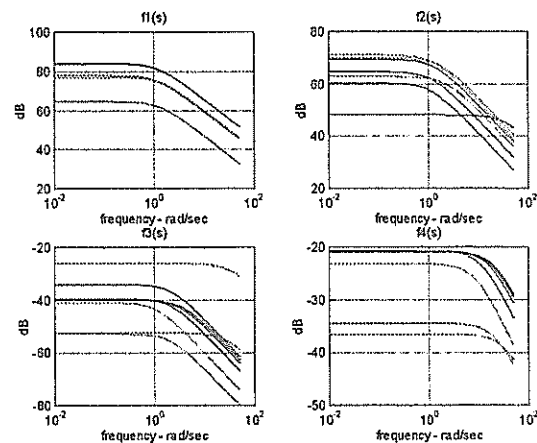


Fig. 3. Open Loop Frequency Response Results describing the plant and its uncertainty

The step response tests were repeated for four other machines and the results are the response of the: (i) speed to the speed control signal, $f_1(s)$ (ii) speed to the square-flux control signal, $f_2(s)$ (iii) square-flux to the speed control signal, $f_3(s)$ (iv) square-flux to the square-flux control signal, $f_4(s)$. Using standard modeling techniques the time domain results were converted to the frequency domain for analysis and design.

The results describe the entire range of machines of concern, hence describing the uncertainty of the responses in full. For ideal decoupling $f_2(s)$ and $f_3(s)$ should have zero response. The uncertainty can now be quantified and used in the design of the feedback controllers.

ROTOR RESISTANCE ESTIMATION

Krishnan's[4] work on the sensitivity of parameters in indirect vector control was reviewed. The slow drifting internal temperature of the machine causes a theoretical linear drift in the resistance which in turn causes a drift in the model away from its nominal.

The energy balance describing the relationship between the stator and rotor has been formulated:

$$P_s - h_{rs}(T_s - T_r) - h_{er}(T_s - T_{env}) = P_r - h_{rs}(T_r - T_s) \quad (2)$$

$$i_s^2 r_1 - (h_{rs} - h_{er} - h_{rs})T_s + h_{er}T_{env} - \alpha_{rs}i_s^2 r_1 = h_{rs}T_r \quad (3)$$

where

i_s : stator current

r_1, r_2 : stator, rotor resistance

P_s, P_r : stator, rotor power

T_s, T_r, T_{env} : stator, rotor and environment temperature

h_{ij} : are transfer coefficients which depend on the machine parameters

Assuming a linear relationship between the resistance and the temperature, one can write

$$r_2(k) - r_2(k-1) = \alpha_{rt} T_r \quad (4)$$

$$i_s^2 r_1 (1 - \alpha_{rs}) - \alpha_{rt} T_s + \alpha_{et} T_o = h_{rs} / \alpha_{rt} (r_2(k) - r_2(k-1)) \quad (5)$$

$$r_2(k) = \alpha_{i_s} i_s^2 - \alpha_{c12} T_s + \alpha_{c21} T_{env} - r_2(k-1) \quad (6)$$

For a small difference in the temperature between the stator and immediate environment eq. 6 becomes:

$$r_2(k) \approx r_2(k-1) + i_s^2 \alpha_{i_s} (+ f_{NL}(T_s, T_{env})) \quad (7)$$

The estimator was tested with forced rotor resistance drifts for the nominal, upper and lower worst case machines in the range. The results are plotted in fig. 4.

For the three cases with the nominal, upper and lower machines, α_{i_s} was found to have values of 0.4, 0.06 and 0.029 respectively. This uncertainty in α_{i_s} was incorporated into the feedback controllers to improve the robustness of the estimator making it universal to all machines in the range.

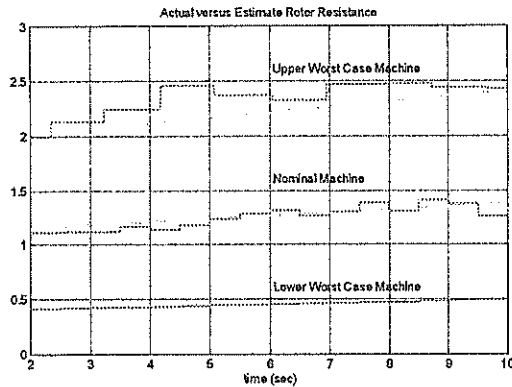


Fig. 4 Actual versus Estimated Rotor Resistance

POWER EFFICIENCY CONTROLLER

The parabolic relationship between the square-flux and the power consumption demonstrated by Kim was verified. The algorithm is intended to simultaneously track and maintain highest power efficiency and perform field weakening. If a load or speed change is detected the flux is immediately set back to the rated machine flux to always ensure highest dynamic performance. From the square-flux - power consumption relationship, the following model has been proposed for the minimum power search algorithm:

$$\phi^*(k) = \phi^*(k-1) + k_{\phi p} \Delta P \quad (8)$$

where $\phi^*(k)$ is the reference square-flux

$k_{\phi p}$: power to square-flux transfer coefficient

ΔP is the difference in power over the last 2 samples

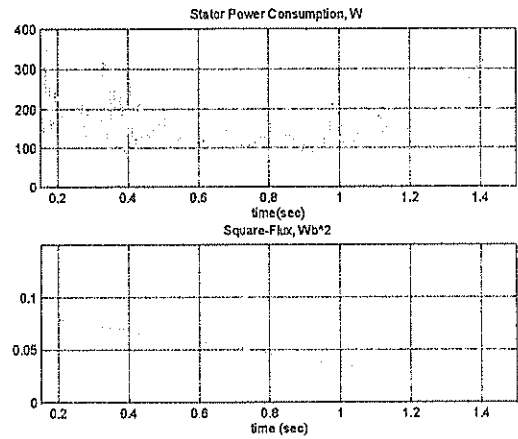


Fig. 5 Power Consumption Results with varying Square-Flux

$k_{\phi p}$ was calculated by varying the square-flux by $90e-05 \text{ Wb}^2$ at each integration step. The tests were carried out for a speed and initial square-flux reference of 3000 rpm and $90e-03 \text{ Wb}^2$ respectively and constant load torque of 0.2 Nm for the nominal machine used. The results are shown in fig. 5.

A parabolic shaped power consumption curve is apparent. The machine power consumption drops from an average value of 0.250 kW to 0.125 kW, for a drop in the square-flux from 0.09 Wb^2 to 0.038 Wb^2 and thereafter increases from approximately 0.125 kW at 0.038 Wb^2 to 0.34 kW at 0.02 Wb^2 .

The results obtained in searching for the optimum operating flux were good. The rotor resistance estimator is also a function of the power consumption and is disabled whilst searching for the optimum flux level.

SPEED AND SQUARE-FLUX FEEDBACK CONTROLLERS

For plants where uncertainty has an unacceptable influence on the output performance, Quantitative Feedback Theory developed by Horowitz[5], offers a design methodology to ensure that the dynamic performance specifications are met. Once the uncertainty in the plant has been quantified it can easily be implemented into the design process, and through a step-by-step process the specifications at the critical frequencies are satisfied.

The speed performance specifications are:

1. Response Time Range: 0.15s to 0.3s; peak overshoot of 10%; and a steady state error of 8%
2. Load Disturbance Specification: $\leq 5\%$ maximum speed change due to a step load torque by 0.5 Nm
3. Coupling Disturbance Specification: $\leq 5\%$ maximum speed change for 0.09 Wb^2 square-flux magnitude change

4. Uncertainty Specification: same as 1-3 above for induction machines ranging from 0.37 to 0.75kW .

The square-flux performance specifications are:

1. Response Time Range: 0.3s to 0.6s; a peak overshoot of 41%; and a steady state error of 8%.
2. Uncertainty Specification: same as 4 above.

The critical frequencies were chosen for each system. Plant templates describing the variations that the plant may assume were calculated at the critical frequencies. The nominal speed and square-flux controllers were chosen; and the speed and square-flux were calculated at the critical frequencies.

On the Nichols chart the boundary conditions including the Universal High Frequency Boundaries i.e. the boundary that ensures the peak overshoot is within limits, were plotted. Using the templates the boundaries were modified to include the uncertainties. The nominal plants were plotted on individual Nichols charts. The open loop term L , which is the product of the plant and feedback controller was used to tune the closed loop response of the speed and square-flux individually.

$B(\omega)$ represents one frequency boundary. The speed and square-flux feedback controllers are:

$$G_{\omega}(s) = \left(\frac{0.02}{s} \right) \times \left(\frac{s / 4.04 + 1}{s / 89.21 + 1} \right) \times \left(\frac{s / 150.93 + 1}{s / 866 + 1} \right) \quad (9)$$

and

$$G_{\phi}(s) = \left(\frac{100}{s} \right) \times \left(\frac{s / 35.38 + 1}{s / 50 + 1} \right) \quad (10)$$

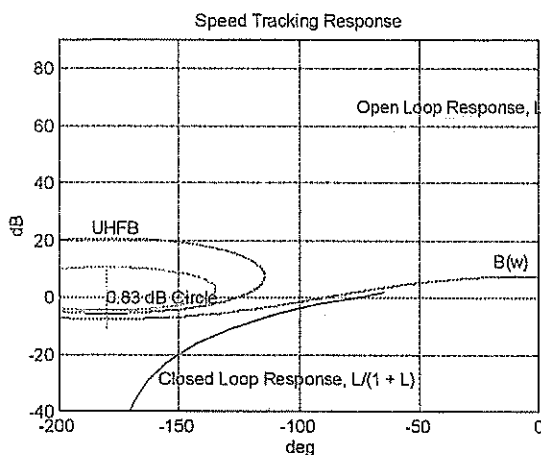


Fig. 6 Nichols Chart of Speed Open versus Closed Loop Response

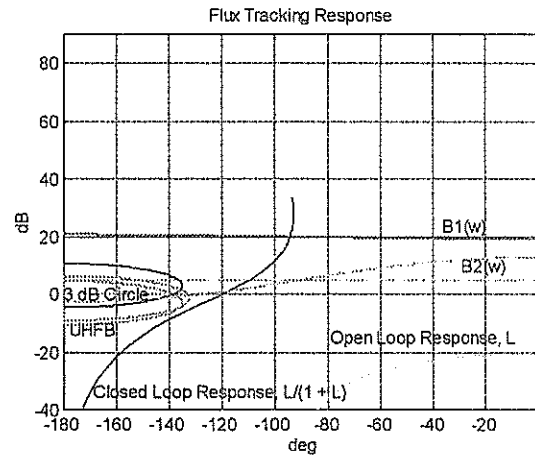


Fig. 7 Nichols Chart of Square-Flux Open versus Closed Loop Response

$B1(\omega)$ and $B2(\omega)$ are also frequency boundaries. A set of results including the upper and lower speed limits, the reference speed and square-flux; and the responses for the nominal, lower and upper worst case plants are shown in fig. 8.

The fixed QFT controller produced remarkable results for all the machines in the design range. The pi controllers produced very good responses for the nominal; the gains had to be changed for the different machines as was expected.

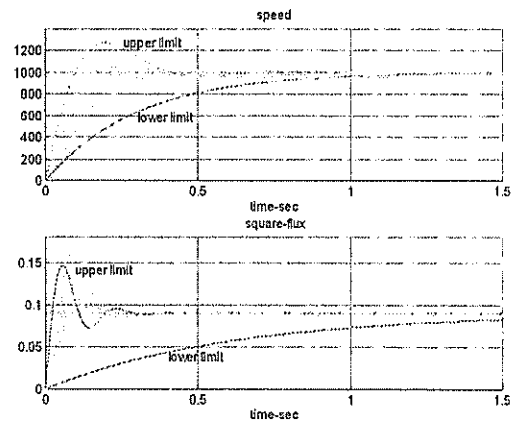


Fig. 8 Speed and Square-Flux Response

CONCLUSION

The paper has shown that by accurately quantifying the uncertainty in a plant or for a range of plants, the feedback controllers responses can be designed to meet specific dynamic performance. Using QFT a speed and square-flux controller were designed with insight of the trade-offs and practical limits. The results are displayed in fig. 8; although final optimisation of the controllers has not yet been done the results are impressive. A rotor resistance estimator based on the change in stator power due to changes in the rotor resistance, reduced the bandwidth required to achieve the desired performance.

The estimator showed a linear relationship between the stator power and rotor resistance for small changes in the resistance. The results shown in fig. 4 are conclusive that the estimator does work well with a large sample time. The power efficiency controller worked well in steady state and produced excellent results. The control system was designed to be universal for a range of machines, which makes it more industrially acceptable.

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DESIGN AND CONSTRUCTION OF A LEVITATED VEHICLE USING A SINGLE EXCITATION SOURCE

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Magnetically levitated (MagLev) vehicles use two or three excitation systems: (1) the propulsion system, (2) the suspension system and (3) the lateral stabilisation system. This project aimed at determining the feasibility of combining all three excitation systems into a single excitation system. It was found that a *Linear Induction Motor* (LIM) with a "V-shaped" long primary design provided the required stabilisation force as well as the required thrust and levitation force. The short secondary consists of a non-ferromagnetic aluminium sheet to form an electrodynamic levitation system. An application of the system has been identified within the aluminium can manufacturing industry. Preliminary test results at the optimal supply frequency of 50 Hz revealed a maximum thrust and levitation force of approximately 85 N and 70 N respectively.

Keywords: linear induction motor, MagLev, electrodynamic

1. INTRODUCTION

Magnetically levitated (MagLev) vehicles are either levitated using an electromagnetic levitation system or an electrodynamic levitation system. In electromagnetic levitation systems the attractive force between a steel yoke and an electromagnet levitates the vehicle, whereas electrodynamic levitation systems use repulsive forces between a super conductive electromagnet, or linear motor, and a non-ferromagnetic aluminium sheet to levitate the vehicle [1]. Normally three excitation systems are present in a MagLev system, namely: (1) the propulsion system, (2) the suspension system and (3) the lateral stabilisation system. All three systems can, however, be combined into a single excitation system. In order to successfully design and construct such a system, the problems associated with the open-ended structure of linear motors need to be considered. This paper will present the design and construction process of such a MagLev vehicle. Analytical equations and *Finite Element Methods* (FEM) will be used to validate design decisions.

2. THEORY OF OPERATION

The electrodynamic magnetic levitation system consists of a long primary with a levitated short secondary. Figure 1 shows that the levitation of the non-ferromagnetic secondary is achieved as a result of eddy currents induced in the aluminium sheet, which sets up a *Magneto Motive Force* (MMF). The secondary MMF interacts with the primary MMF produced by the primary coils, to create a repulsive (levitation) force. The thickness of the aluminium secondary determines the resistance of the aluminium and the interaction between the MMFs produced by the primary and the secondary.

A fundamental difference between a rotary induction motor and a LIM is the finite length of the magnetic and electric circuit of the LIM in the direction of the

travelling magnetic field. The open ended magnetic circuit produces a longitudinal end effect, as well as the so-called transverse edge effect.

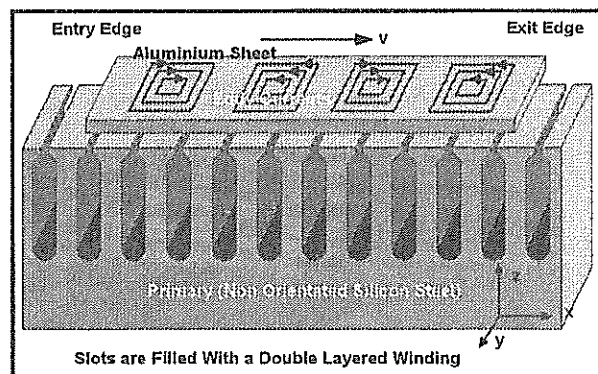


Figure 1: An electrodynamic levitation system

2.1 Longitudinal End Effect

The end effect can be divided into the static end effect and the dynamic end effect [2]. The open-ended primary magnetic circuit creates a so-called static longitudinal end effect. This is a result of the asymmetrical placement of the three phase coils with respect to the entry and exit ends. The travelling magnetic wave is effectively "cut off" at each end of the LIM. Interference between the synchronous travelling wave and the reflected waves at each end of the LIM results in the superposition of the waves to form a standing wave. The normal component of the air gap flux density is thus weakened at the entry edge and amplified at the exit edge [1]. This characteristic causes the secondary to be levitated at an angle with respect to the primary.

The dynamic longitudinal end effect is dependent on the steady-state velocity of the secondary. An increase in the velocity of the secondary will result in the longitudinal end effect being more prominent. Dynamic longitudinal end effects for high speed

LIMs causes a noticeable reduction in the efficiency and power factor in the low slip region [2].

Longitudinal end effects are characterised by [1]:

- A non-uniform flux density distribution in the air gap.
- A non-uniform eddy current distribution in the secondary.
- Unbalanced phase currents.
- Parasitical braking forces.

The higher the slip at which the LIM is operated the smaller the longitudinal end effect will be. The global influence of the longitudinal end effect decreases with an increase in the number of poles of the primary.

2.2 Transverse Edge Effects

Transverse edge effects appear only in flat LIMs with a primary and secondary of finite width. The transverse edge effect is characterised by [1]:

- A non-uniform magnetic flux density distribution in the air gap along the y-axis. The flux density increases at the edges of the primary due an increase in the reluctance and the eddy current reaction in the secondary.
- A generation of a destabilising lateral forces perpendicular to the travelling MMF wave.

The effect will decrease with an increase in the ratio of secondary width to primary width, up to the optimal point.

2.3 Open Primary Slots Versus Semi-Closed Primary Slots

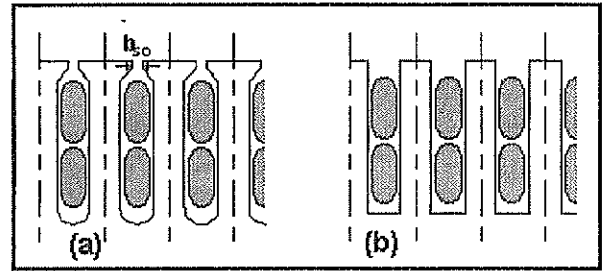
LIM primaries can be designed with open or semi-closed slots. Figure 2(a) shows the new semi-closed slot design and figure 2(b) shows the open slot design proposed by Laithewaite [3], which have been previously used for linear motor research at the University of the Witwatersrand. Open slot geometries allow pre-formed coils to be inserted into the slots. A disadvantage of using the open slot geometry is the small air gap that must be maintained (approximately 1mm or less) in order to obtain a large air gap flux density [4].

The Carter coefficient for open slot geometries is large compared to semi-closed slot geometries. The Carter coefficient is given by:

$$K_c = \frac{1}{1 - \gamma g / \tau} \quad (2.3.1) \quad \dots$$

where γ can be approximated as [2]:

$$\gamma \approx \frac{(b_{so} / g)^2}{5 + b_{so} / g} \quad (2.3.2) \quad \dots$$



where b_{so} = slot opening, g = air gap distance

Figure 2: Slot geometries: (a) semi-closed and (b) open.

Although semi-closed primary slots are more costly and difficult to manufacture, air gap distances between primaries and secondaries can be increased, depending on the required flux density in the air gap. A trade-off exists, however, where increasing the air gap distance for a semi-closed slot will result in an increase in the leakage permeance. For open slots the leakage permeance is decreased for high air gaps of 10 to 15 mm [2]. Since the vehicle is to be operated at small air gaps (3 to 5 mm) a better flux density can be obtained in the air gap with a semi-closed slot than with the open slot. The aluminium sheet secondary will have to be loaded, however, to maintain the required air gap.

3. THE MAGLEV SYSTEM

A primary design proposed by Laithewaite [3] was used as a basis from which a new and improved design could be developed. The striking differences between Laithwaite's [3] design and the design shown in figure 3, is the inclusion of an extra pole, as well as the use of a semi-closed slot geometry.

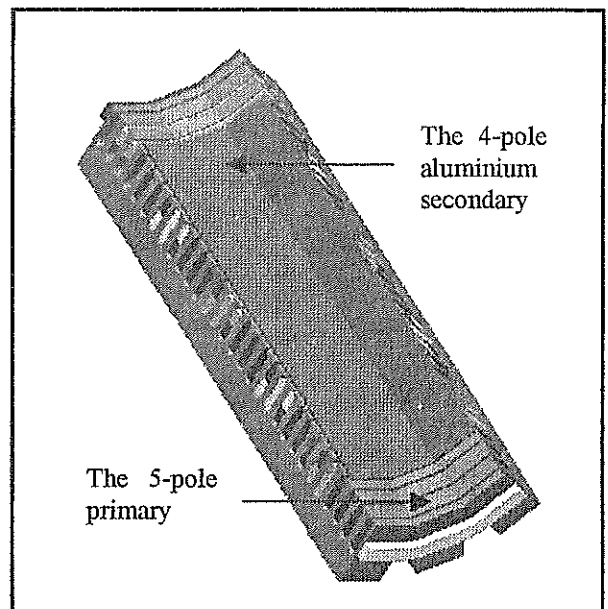


Figure 3: The MagLev system

3.1 Primary Design

The primary design is based on the basic structure of a longitudinal tubular linear induction motor (TLIM) cut in half. TLIMs have a closed geometry about the radial axis, such that no transverse edge effects can be produced. This is also the philosophy behind the use of the “v-shape” arrangement of the primary sections. The two edge primary sections are inclined at 35 degrees. The original design specified that the sections be inclined at 30 degrees, but was changed due to construction restraints. The two inclined primary sections provide a stabilising component as well as a levitation component. The “v-shape” arrangement produces a balancing force, which in turn reduces the effect of the transverse edge effects. At this point it is worth mentioning the choice of the odd pole LIM. Studies have shown that for small speeds, the even pole primary’s performance is slightly inferior to that of the odd pole [2]. Table 1 gives the primary design specifications.

Table 1: Primary design specifications.

Description	Dimensions
Number of slots	30
Pole pitch	114 mm
Slot width	10 mm
Total slot depth	42 mm
Effective slot depth	36.5 mm
Slot opening	3 mm
Bottom slot radius	5 mm
Total tooth width	16 mm
Effective tooth width	9 mm
Primary width	50 mm
Primary height	81 mm
Primary length	570 mm

The average flux density, B_{air} , in the air gap is a function of the primary width, w_p :

$$B_{air} = \frac{\Phi}{3 \times w_p \times \tau} \quad \dots \quad (3.1.1)$$

where Φ = flux per pole.

The width of the primary was chosen such, as to give the required flux density in the air gap, without saturating the stator teeth, as well as to ensure that the inverter could supply the required magnetising current.

3.2 Primary Winding

The primary has a double layered three-phase winding, short corded by one slot pitch. The primary winding is wound from two strands of 0.9 mm diameter varnished copper wire. The choice of the diameter wire was based on a recommendation by

Gieras [1], which states that the current density in primary windings are between 4 and 25 A/mm², depending on the output power and the cooling system. If the primary current density exceeds 15 A/mm², conductors must have internal channels and water-cooling systems [1]. The two strands of 0.9 mm wire therefore has an effective diameter of 1.8 mm and if supplied with a maximum rated current of 16A, a current density of 5.66 A/mm² is achieved. The use of two strands of 0.9 mm wire in stead of one strand of 1.8 mm diameter wire, also facilitates easy insertion of the coils through the small semi-closed slot openings.

Each phase’s coils are connected in series. According to Boldea and Nasar [2], studies have shown that for LIMs up to 1800 kW and 110 m/s, the double-layer fractional-pitch three-phase winding with a total number of odd poles, and half end slots is economically superior to other choices. The winding specifications are given in table 2.

Description	Dimension
Number of poles	5
Number of slots/pole/phase	2
Number of coils/primary	27
Number half filled slots	8
Number of turns/coil	45
Chording factor	5/6
Winding factor	0.933
Maximum rated current	16 A
Normal operating current	14 A
Input voltage	220 V

Table 2: Primary winding specifications

To combat the effect of longitudinal end effects, compensating coils can be included [1]. This principle is applied between two primary track sections. Overlapping coils between the two primary sections ensure a uniform distribution of the MMF wave along the x-axis. Figure 4 shows the overlapping coils.

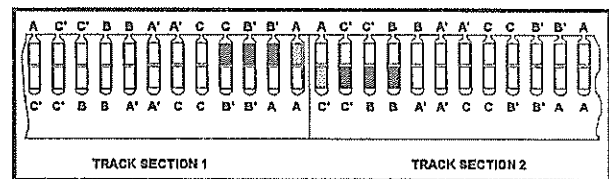


Figure 4: Compensating overlapping coils between two primary sections. Extra coils added are shown in red and existing coils are shown in grey.

4. PERFORMANCE

4.1 Finite Element Analysis

Finite element simulations, using Ansoft's Maxwell 2D and 3D Finite Element Package, were used to determine the effectiveness of the primary design. Since no back iron is used in the secondary, it was necessary to obtain a plot of the field distribution in the air gap and the aluminium secondary. Of particular interest is the leakage permeance, since too much leakage across the slots reduces the levitation and thrust forces. The added fifth pole, which provides an added path for the leakage flux to flow in, reduces the overall leakage flux.

4.2 Performance Measurement

The optimal operating frequency of the vehicle is obtained at 50 Hz for the Varispeed 606PB3B inverter. A maximum thrust and levitation force of 85 N and 70 N respectively, is obtained. An increase in the supply frequency results in a drop in these forces.

5. APPLICATIONS

Crown Cork, a company situated in Roodekop, South Africa, has expressed a keen interest in the possible application of the system for the high-speed transport of aluminium cans in the production line. The company is involved in the production of steel and aluminium cans for the soft drink and beverage industry. The current production line uses conveyer belts and pneumatic levitation systems to transport the cans at high speed through the production process. Guide rails are attached to the side of the conveyer belt to prevent the cans from falling off. Due to the high quality standards employed by the company any small scratches and dents detected on cans during the manufacturing process is rejected. This leads to losses, which with the current increase in the price of materials, adds up to substantial unnecessary costs in the long run. The pneumatic system is only employed in certain small areas of the plant and levitates the cans while transporting them. The thrust force in the pneumatic system is obtained by displacing the holes between two layers of metal sheet. This generates a normal as well as a horizontal component of force. Guide rails are however still used which can cause damage to the cans.

It is proposed that the MagLev system can do the same function, by levitating the cans. The advantage of this MagLev system, however, is that lateral stabilisation forces are already provided so that no guide rails are required. It is expected that by increasing the size of the system large volumes of cans be transported at the same time, similar to the existing conveyor system. The disadvantage of the system is the initial cost of installation, which will be

substantial compared to pneumatic and conveyer belt systems. The advantage of the system is that there are no moving parts, which cuts down on wear and tear and preventative and forced maintenance of the system. The life expectancy of the MagLev system is also longer compared to the pneumatic and conveyer belt system.

6. CONCLUSION

It was found that the levitation, thrust and lateral stabilisation forces for a MagLev system can be produced from a single excitation source. Although the system is inefficient in terms of power usage due to a poor power factor and high slip, the advantage of such a system far out ways the disadvantages. The application of the system as a substitute for current transportation systems should therefore be investigated.

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DESIGN OF A SIMPLE PI CONTROLLER FOR THE REDUCTION OF THE EFFECT OF DETENT FORCES IN A PERMANENT MAGNET LINEAR SYNCHRONOUS MOTOR

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The Detent forces of a permanent magnet linear synchronous motor cause a oscillatory motion by introducing a sinusoidal disturbance. A simple PI controller with velocity feedback was designed. The velocity was measured from the position fed back by an optical encoder at 500Hz. Test showed that the proportional term (K_p) reduced the fluctuations in the error between the measured and the ideal reference velocity, but not significantly. By increasing the integral term (K_i) the error was driven to fluctuate around zero, closer to the desired reference value. An optimum combination of K_p and K_i were found to be 45 and 100 respectively. The resulting motion was substantially smoother and the measured velocity followed the reference value more closely. But physical deficiencies and an unresponsive control strategy limited the performance of the system. Further improvement would be achieved by using a surface mounted permanent magnet linear synchronous motor with semi-closed slots.

Keywords: Buried Permanent Magnet Linear Synchronous Motor, Detent Forces, PI controller

1 INTRODUCTION

Magnetic levitation using **linear synchronous motors (LSM's)** has opened the way to a vast range of applications in the transport of people and goods. The possible implementation of ropeless elevators in ultra-deep shafts has raised the interest of the mining industry. The Machine Research Group at the University of the Witwatersrand has constructed a small prototype of a LSM in order to investigate its behaviour through computer simulation and physical measurement.

The thrust force of the **permanent linear synchronous motor (PMLSM)** consists of three components, namely the main synchronous force, the reluctance force and the Detent force. Both the synchronous and the reluctance force are function of the load angle and are thus constant for a constant angle. The Detent forces arise from the interaction between the permanent magnets and the slotted nature of the stator [1]. They are in fact a function of the position of the mover relative to the stator. It was found that the effect of Detent forces was the superposition of a considerable periodic disturbance in the motion of the PMLSM. This resulted in an oscillatory motion of the mover, which in turn was unacceptable for applications in the mining industry.

The effect of the Detent forces can be reduced by optimising the geometry of the machine [2] and by implementing a control strategy to ensure a continuous motion. In order to ascertain the effectiveness of the proposed control strategy, an existing PMLSM with an unoptimised design was used.

2 BACKGROUND

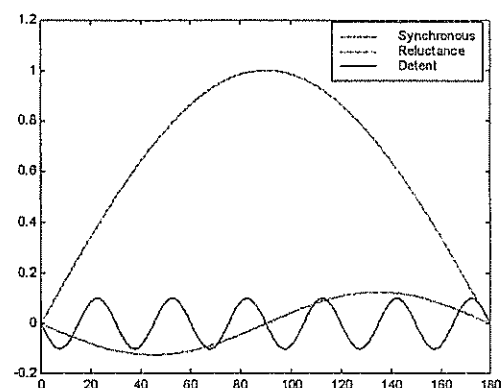


Figure 1: The three normalised components of the static thrust force

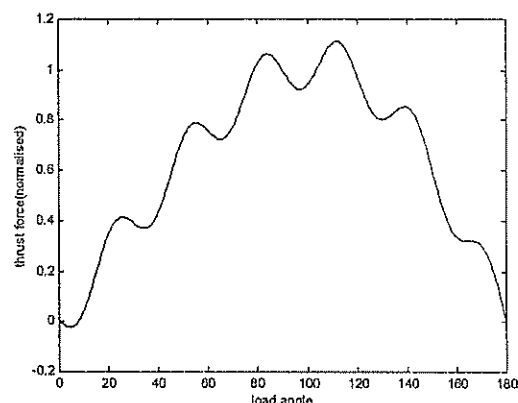


Figure 2: The resultant predicted thrust force

A theoretical representation of the three force components and their resultants is shown in Figure 1 and Figure 2. Since the prototype has 6 slots per pole pitch a sinusoidal distribution of the Detent force with 6 peaks per pole was predicted. The resultant static

thrust force shows that the translator is expected to have an oscillatory or jerky motion.

3 THE BASIC CONTROLLER

The prototype used for this project was designed by Le Rouge and El-Hage [3]. The long primary was an open slot system with four stator sections, 6 slots per pole.

The mover consisted of five buried **permanent magnets (PM)** interspersed by four mild steel pole shoes. The PM material was **Neodymium-iron-boron (NdFeB30)** with a residual flux density of approximately 1.25 T and a coercive force of 10.5 kA/m. A quadrature optical encoder was mounted on the mover to feedback the position.

A FEM prediction of the static thrust force for the prototype was done in a previous project [4], which compared well with the measured force of 600N produced at approximately 90°-load angle.

A Pascal program, originally written by Le Rouge and El Hage [3] controlled the motion of the mover. According to the inputs for the desired maximum velocity and the final destination that the mover must attain, an ideal velocity vs. position curve was computed, as depicted in Figure 3.

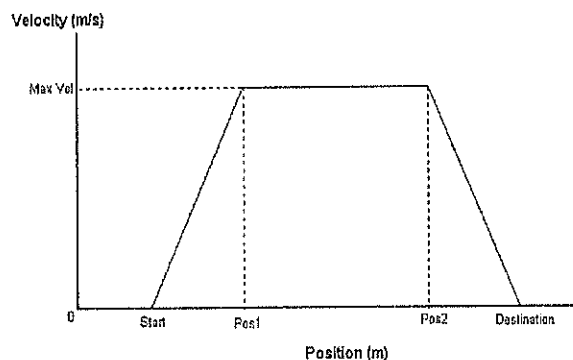


Figure 3: Ideal Velocity vs. Position graph

In essence the motion had three stages. In Stage 1 the mover was first accelerated from the starting position to pos1 where the desired maximum velocity was reached and then maintained during Stage 2. Finally, once pos2 was reached, Stage 3 started where the mover decelerated to a standstill. Due to the disturbance in the motion a tolerance of 0.05m was allowed when triggering the third stage.

During operation, the optical encoder fed back the position of the mover to the PC. A basic controller used the reference ideal velocity vs. position graph to calculate the corresponding velocity and then converted it to its equivalent reference frequency value. Then this value was sent to an inverter, which interfaced the PC to the mover through volt/hertz

control. The controller performed this operation with a sampling frequency of 500Hz.

3.1 Results

The motion of the mover was observed over three stator sections that corresponded to a vertical displacement of 0.8m. The position read from the optical encoder was also sent to a file continuously at 500Hz. A Matlab™ program then plotted the actual outputs as well as the ideal graphs (Figure 4). All the measurements were derived from the same set of inputs: a maximum velocity of 0.4m/s, a displacement of 0.8m.

The results are best shown on the position vs. time graph, plotted in Figure 4. As was expected, the motion was very jerky because of the Detent forces and the gap between the stator sections. The Detent forces gave rise to a sinusoidal disturbance in the motion, as assumed, with at least 15 distinct peaks.

The time delay resulted from the oscillatory motion and the fact there was no time reference employed in the basic control strategy.

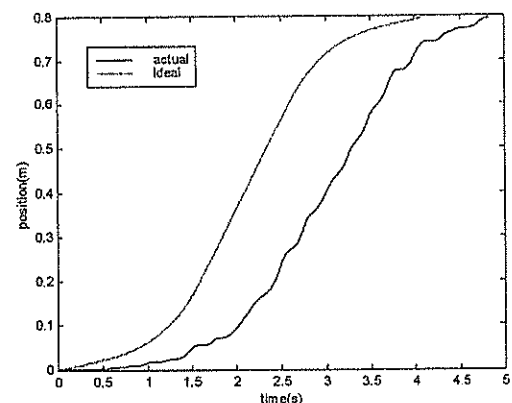


Figure 4: Position of Measured vs. Ideal output

Through Matlab the velocity output was derived by differentiating the position inputs at 500Hz. Figure 5 depicts the oscillatory motion, but by comparison with the ideal velocity vs. position curve, the mover had a slower acceleration in Stage 1, consequently a shorter Stage 2 and was practically stopped to standstill in Stage 3.

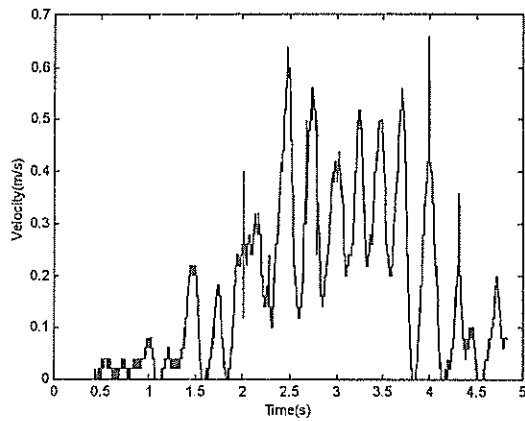


Figure 5: Measured velocity output

4 IMPLEMENTATION OF A SIMPLE PI CONTROLLER

In order to reduce the oscillatory motion a simple digital PI controller with velocity feedback was implemented. The general form for the output (U_{PID}) of a proportional-integral-derivative (PID) controller is [5]:

$$U_{PID} = K_p e + K_D \frac{de}{dt} + K_I \int e(t) dt \quad (1)$$

where K_D , K_P and K_I are the constants for the derivative, proportional and integral controllers. The output U_{PID} is a function of the error between the set output and the measured output. The derivative controller was omitted because of the fast and large fluctuations observed in the measured output. The digital controller was of the following form [5]:

$$V(n) = V(n-1) + (K_p + K_i T) \cdot E(n) - K_p \cdot E(n-1) \quad (2)$$

where

- $V(n)$ = New Velocity output
- $V(n-1)$ = Previous Velocity output
- K_p = Proportional gain
- K_i = Integral gain
- T = Sample period, 2ms
- $E(n)$ = Present Error
- $E(n-1)$ = Previous Error

$V(n)$ represents the average velocity of the mover calculated over 5 successive position inputs, to be subtracted from the ideal reference value in order to deduce the error.

Stage 2 and Stage 3 were triggered automatically once the mover had reached pos1 and pos2 respectively. That improved the synchronisation between the motion of the mover and the stages and ensured that the mover

never exceeded the 0.8m travel distance. The block diagram of the controller was as follows:

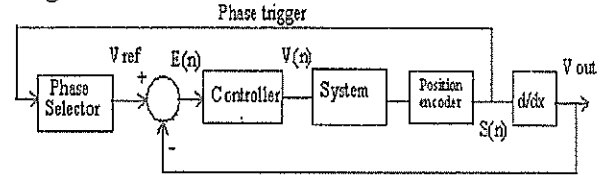


Figure 6: Block diagram of Controller

Adding an additional PI controller with position as the setpoint did not improve the results, and thus was omitted.

4.1 Results

Various tests were run with different values of K_p and K_i . The position, the average velocity and the calculated error were sent to a text file that was then processed by Matlab™.

The results have shown that the proportional controller (K_p) does not effectively attenuate the fluctuations in the error between successive peaks and troughs to an overall steady state. But the integral controller (K_i) successfully drove the overall steady state of the system closer to the desired set point. In Figure 8 for $K_i = 20$ the error fluctuated around 0.4 m/s, while in Figure 7 where $K_i = 100$ the oscillations occur around 0 m/s, suggesting a closer attempt to settle down at the desired set point.

The optimum combination of K_p and K_i were found to be 45 and 100 respectively for it produced the smoothest motion with all three stages distinctly evident. The sinusoidal disturbance in the motion was considerably reduced (Figure 10). Also the velocity vs. time followed more closely the trapezoidal shape of the ideal graph (Figure 11).

Greater values of K_p and K_i produced excessive accelerations and rendered the system unstable at times.

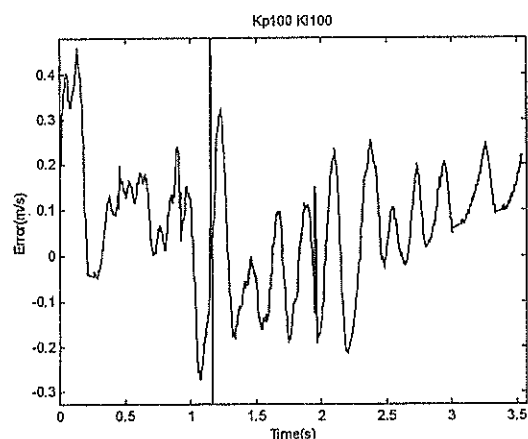


Figure 7: Velocity error vs. time for $K_p=100$ $K_i=100$

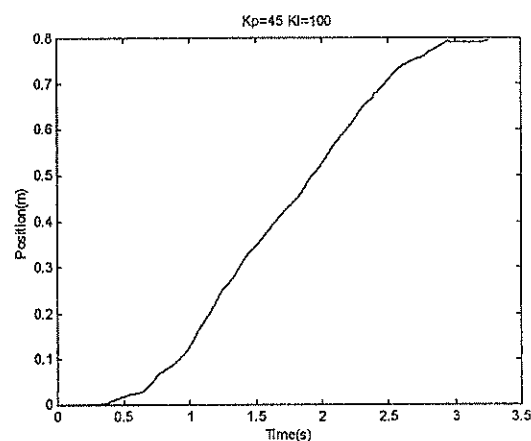


Figure 10: Improvement in Motion with PI controller

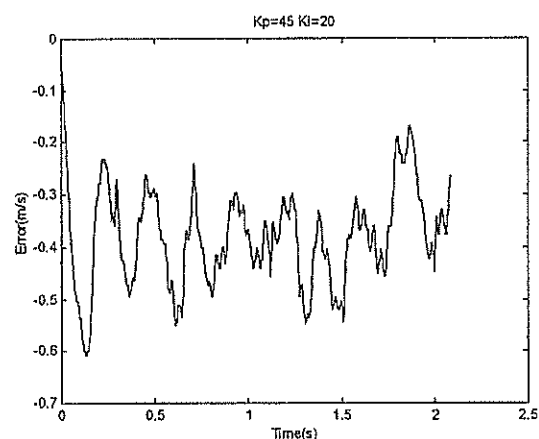


Figure 8: Velocity error vs. time for $K_p=45$ $K_i=20$

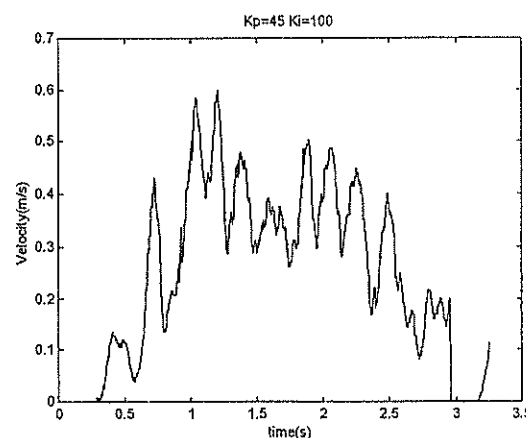


Figure 11: Velocity output with PI controller

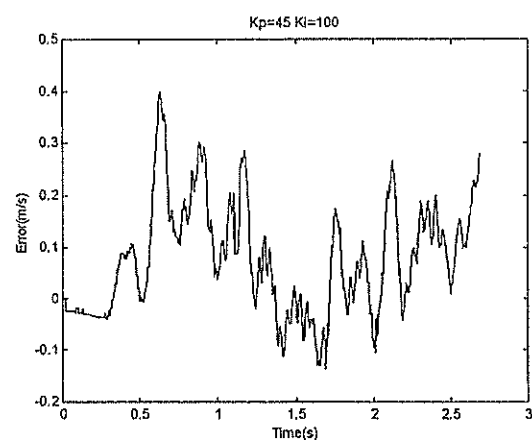


Figure 9: Velocity error vs. time for $K_p=45$ $K_i=100$

5 PERFORMANCE LIMITATIONS

The performance of the PI controller was limited by the system's physical deficiencies. The gap between the primary and the secondary was irregular during the mover's travel. It was deviating from the set value of 3 mm because of shifts in the stator sections and difficulties in aligning the mover. These deviations caused an imbalance in the axial forces resulting in the mover being pulled closer to either side of the stator sections, scraping the side at times.

The control strategy implemented was based on scalar control which does not allow for the direct control of the driving forces (i.e. synchronous and reluctance).

Driven by the error between the measured and the reference value, the PI controller is not proactive. It continuously attempts to rectify an existing error. Furthermore time constraints prevented a thorough modelling of the system.

6 CONCLUSIONS

The Detent forces in a permanent linear synchronous motor arise out of the interaction between the slotted nature of the secondary and the ends of the permanent magnets on the primary. This resulted in the superposition of an overall sinusoidal disturbance in the motion of the mover.

The motion of the mover had 3 stages each triggered by the position fed back by the optical position encoder. In order to reduce the oscillatory motion a PI controller with velocity feedback was implemented. The motion of the mover was substantially improved. The resultant position vs. time graph became smoother. The measured velocity output followed the reference ideal graph much more closely. The proportional term (K_p) did not improve significantly the steady state response of the system. But the integral term (K_i) successfully drove the error in the velocity output to fluctuate around zero m/s, closer to the desired reference value. No derivative term was implemented because of the large and rapid fluctuations in the error. Without a filter, a derivative controller would have produced an undesirable response.

The performance of the PI controller was limited by physical deficiencies of the system that added additional disturbance to the motion. The inability of the controller to either directly control the driving forces or to be proactive limited its performance.

Hence for further improvement an alternative control strategy must be considered which will also take into account a thorough modelling of the system.

But in essence, the performance would be optimised by rebuilding the system to eliminate the physical shortfalls. The Detent forces would be considerably reduced by the use of:

- semi-closed slots [2]
- skewing of the magnets or primary sections [2]
- a surface mounted permanent magnet permanent linear synchronous motor with optimised magnet sizes [4].

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HARD REAL-TIME CODE GENERATION USING SIMULINK AND REAL-TIME WORKSHOP

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Abstract - Simulink and the Real-Time Workshop (RTW) from Math Works allow automated generation of executable code from block diagrams. This paper illustrates how to ensure that the generated code meets hard real-time deadlines as necessary for motion control applications. Specific emphasis is placed on generating code to execute predictably in an asynchronous interrupt driven environment

Index - code generation, motion control, rapid prototyping, real-time, RTW, Simulink

1 INTRODUCTION

A real-time system is : "Any system in which the time at which output is produced is significant. This is usually because the input corresponds to some movement in the physical world, and the output has to relate to the same movement. The lag from input time to output time must be sufficiently small for acceptable timeliness."

Oxford Dictionary of Computing [1]

It can be argued that all practical systems are real-time in nature, i.e. they have to respond within some time constraint. Batch oriented systems, such as those operated by insurance companies or magazine subscriptions may have response times of several weeks or more and still be considered satisfactory. Thus, batch oriented systems are seldom referred to as real-time. On line or interactive systems, such as word processing programs, should respond to commands within a reasonable amount of time or they become very inconvenient to use. Most literature refers to these systems as *soft real-time* systems as their performance is degraded but not completely destroyed by failure to meet response time deadlines. *Hard real-time* systems are generally characterised by direct interaction with the environment, and by strict time constraints, whose violation might have severe consequences, Fig.1. Examples of hard real-time systems include flight control, telecommunication, motion control etc. [2] Generally most such control loops operate within very narrow time margins. Failure to remain within those

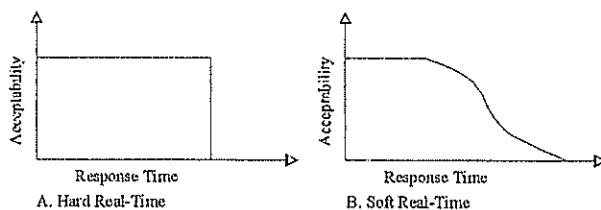


Fig. 1. Difference between Hard (A) and Soft (B) real-time systems

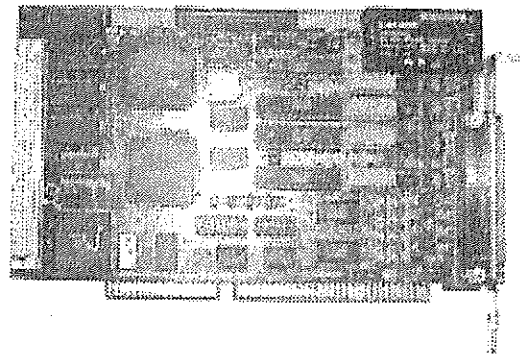


Fig. 2. Photo of the Innovative Integration PC32 card.

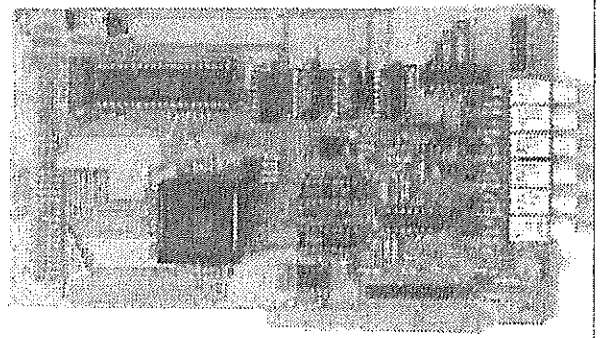


Fig. 3. Photo of the custom PWM / Tacho card.

constraints means loss of control and could be catastrophic.

The paper concentrates on a system which automatically generates hard real-time interrupt based code for motion control applications from block diagrams. The system is based around the well known Matlab, Simulink and Real Time Workshop (RTW) environment from Math Works Inc. [3, 4, 5]. Hardware targeted is the PC32 card from Innovative Integration, Fig.2, as well as a custom PWM/Tacho card, Fig.3 [6, 7].

In particular a Field Oriented Control algorithm is implemented as a benchmark to validate the performance of the system.

2 PC32 AND PWM CARDS

The PC32, shown in Fig.2, features the Texas Instruments TMS320C32 32-bit floating point DSP capable of 60 MFLOPs/30 MIPs. On-chip peripherals include two 32-bit counter/timers, two flexible DMA controllers as well as 15 prioritized interrupts. Memory on the PC32 may be expanded up to 512Kx32 zero wait and 1Mbyte, 1 wait-state. Included on the card also are four channels of 16-bit, instrumentation-grade analog-to-digital converters capable of sampling up to 100K samples/sec. Also included are four channels of 16-bit, instrumentation-grade digital-to-analog converters at 100kHz per channel. Sixteen bits of high-drive digital I/O are also on-card. The onboard devices can be driven either externally, by the onboard timers or using memory mapped access. The PC32 is a half-card which plugs into a standard 16-bit ISA bus slot. The ISA bus interface provides high-performance dual-ported memory capable of 5 Mbytes/sec. on most platforms. The 4K byte dual port provides an interface to the PC.

The custom PWM/Tacho card, shown in Fig. 3, is used to take the burden of driving the power electronics away from the TMS320C32 processor. Also the optical fibre interface isolates the PC from the power electronics. The card was developed in-house at the Motion Control Group [6, 7].

3 THE REAL-TIME WORKSHOP (RTW)

Matlab, Simulink, RTW and additional toolboxes form an extensive set of tools for modelling, control design, analysis and simulation. Using these tools the desired controller can easily be designed, configured and simulated on a PC. Because of its inherent power, the Simulink environment is becoming widely used in academic institutions and is becoming a defacto standard in terms of simulation.

After a system has been designed and simulated in Simulink, the resultant block diagram can be automatically compiled, linked with the necessary components and libraries and then downloaded onto the dedicated target hardware to be executed in real time.

While the code is executing in real time various parameters can be changed in the Simulink block diagram and may be automatically updated in the code executing on the hardware without interrupting execution [8].

4 GENERATED PROGRAM

As a default code generated by RTW is targeted at a single-tasking environment and implements a scheme to emulate multitasking - pseudo-multitasking. An interrupt routine is tied to a timer and executes at the base sampling rate of the block diagram, all slower blocks need to have a sampling rate that is a multiple of the base sampling rate. This configuration is adequate for most *soft real-time* applications and systems which do not need to be synchronised to the external hardware. Most motion control applications, however, require the control software to respond to asynchronous events.

Thus, the basic functionality of RTW needed to be extended to support multiple asynchronous interrupts.

5 DEVICE DRIVER BLOCKS

In order for the code generated from a Simulink diagram to be able to access low level devices like A/Ds, D/A's and for example the PWM/Tacho card, additional device driver blocks have to be defined and are implemented as S-functions on the block diagram, Fig. 4. When the RTW is invoked the code associated with those

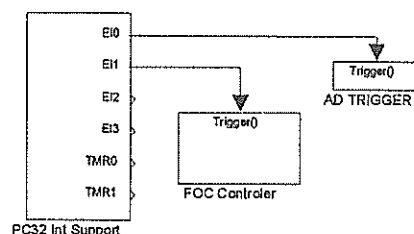


Fig. 4. Top Layer of FOC block diagram

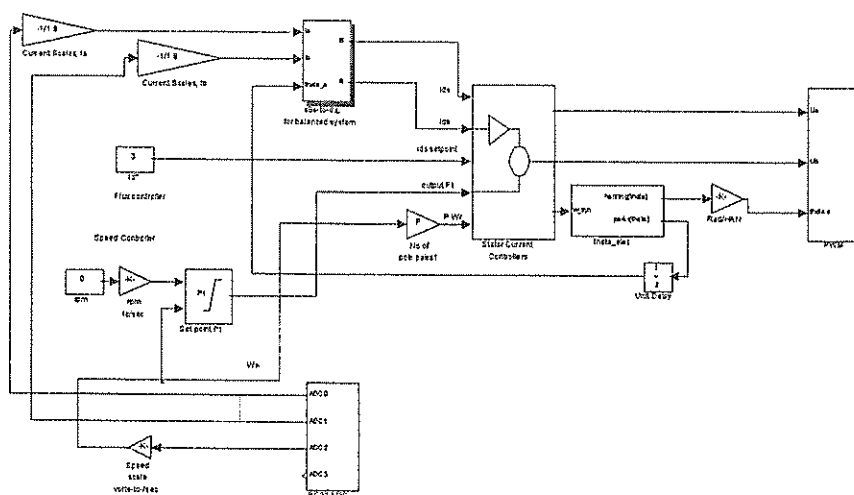


Fig. 5. Field Oriented Control block diagram

S-Functions is automatically compiled and linked into the final code. In addition to providing the basic I/O functionality, S-functions can be used to provide access to custom external hardware.

Fig. 4 shows the basic blocks needed to construct an asynchronous system. The interrupt support block ensures that all downstream blocks or subsystems are installed as separate interrupt handlers. In this particular example an FOC algorithm is attached to external interrupt 2 - which is the end of conversion interrupt generated by the onboard AD. The diagram for the controller is contained in the FOC Controller block in Fig.4. and is shown in Fig. 5.

6 FOC IMPLEMENTATION

Fig. 5 demonstrates a typical setup for controlling an induction machine using field oriented control (FOC). The complexity of the controller becomes evident when one consider that the diagram has another two layers under some of the subsystem blocks Fig. 6, 7. In practice

FOC is very computationally demanding and in fact was impractical to implement until the advent of powerful microprocessors.

The task of writing dedicated FOC code for an embedded controller is daunting even when high level language like C is used. Each small change to the layout of the controller could mean a near complete re-write of the code.

With the Integrated Development System [8], all the hassle of code writing and compiling is done transparently for the user. The development cycle is shortened drastically, so that numerous approaches can be tried out in a short space of time. This means that even users with no or limited experience in programming can develop reliable embedded code.

Fig. 8 shows the speed response of the untuned FOC controlling a 1,5kW induction machine [9]. Fig. 9 shows the stator current I_q and the speed. It can be seen that the stator current drops off after full speed has been reached. This means that the torque of the machine is proportional to the stator current, a characteristic of FOC.

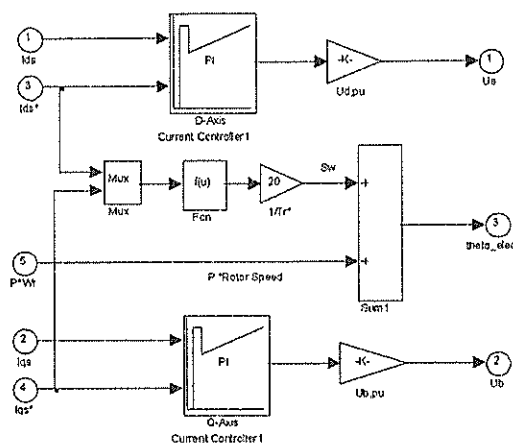


Fig. 6. Stator Currents Controller subsystem

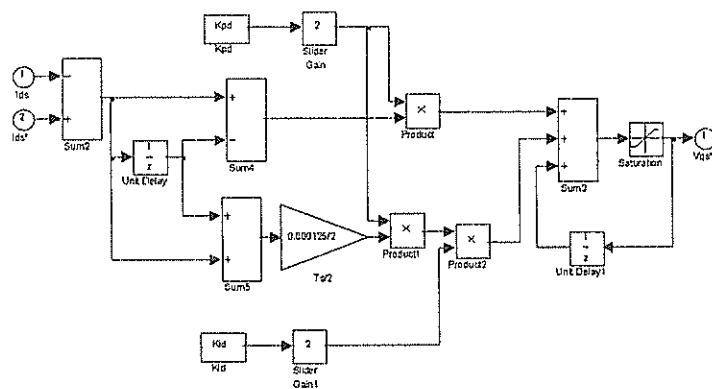


Fig. 7. D-Axis Current Controller subsystem

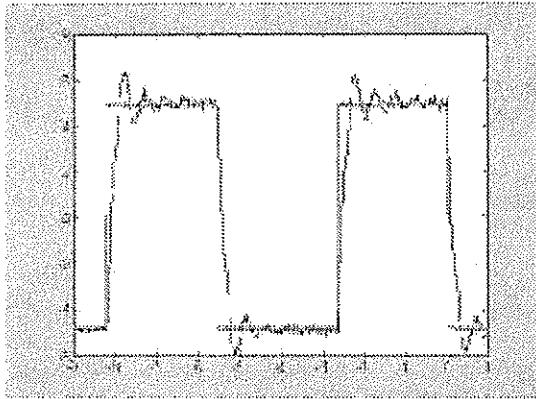


Fig. 8. Speed reference and actual untuned speed response.

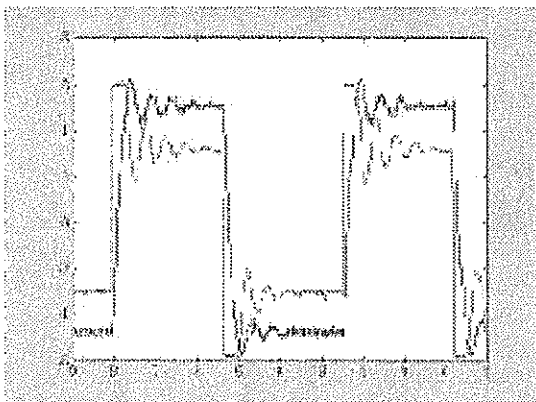


Fig. 9. Speed and Iq plots.

7 CONCLUSION

The system provides a robust tool for any control engineer in need of generating reliable target code which supports asynchronous events. Also it provides a convenient and inexpensive platform for undergraduate students to experiment with complex control options without the need for writing dedicated software. It combines the proven and widely accepted mathematical power of the Matlab environment with the ease of use and hassle free code generation of dedicated packages like Hypersignal.

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AUTOMATED PATTERN RECOGNITION TECHNIQUES APPLIED TO PD MONITORING IN LARGE ELECTRICAL MACHINES

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Abstract: The modern plant is dependent on the proper maintenance of all machinery for optimum production [11]. Condition monitoring involves the measurement and interpretation of large data sets for potentially non-linear relationships. The aim is to enable the accurate and timely prediction of possible malfunctions and the extent of the damage. The measuring of PD pulses in motors, generators and transformers constitutes a specific condition monitoring technique. In practice more than one failure mechanism can occur simultaneously, complicating the diagnosis. While digital processing of PD pulses have enabled automated processing of PD measurements, quantification of the insulation condition remains elusive. Recent research highlights the importance of pulse-shape partial discharge analysis in addition to partial discharge height distribution [3].

Keywords: Partial discharge analysis, automated processing of PD, condition monitoring.

1 INTRODUCTION

Abbreviations:

PCA - Principal component analysis
PD - Partial discharge
FFNN - Feedforward neural network
FLNN - Fuzzy logic neural network (neuro-fuzzy networks)
PDHD - Partial discharge height distribution
PSPDA - Pulse-shape partial discharge analysis

1.1 BACKGROUND

The degradation of insulators in high voltage equipment leads to increased field strength in the insulator, eventually causing a complete breakdown. The resulting PD pulses can be detected using a variety of transducers (acoustic, light detection, RF, $\tan \delta$ etc.) [10].

The early detection of degradation in the insulation can help to reduce potential damage associated with a catastrophic failure. The failure can even lead to the destruction of a plant [11]. Due to the shortage of skilled on site personnel, much effort has been focussed on the automated recognition of PD measurements [6]. The aim is to develop suitable pre-processing techniques to enable classification of PD data by an automated system. Although positive results have been reported, the condition of the insulation is not quantified [6]. This makes it difficult to design a maintenance program that can identify and correct potential problems before they develop. Furthermore, the generalisation performance of the classifier on other machinery is not known.

Traditionally the focus has been on pulse amplitude and pulse phase measurements to quantify PD activity in machinery. Recently high speed digital waveform recorders up to 1 GHz have enabled the recording and analysing of the exact PD waveform, after propagating through the system. The Weibull model can be used to model the shape of the PD pulses as well [1]. The result of laboratory ageing tests indicates that the

ageing process can be modelled by a Weibull model with suitable parameters.

The sparsity of available PD measurements on creates a need to supplement the small data set with data derived from a model of the machinery. An accurate model of the PD mechanism as well as the propagation path is necessary in order to interpret PD measurements meaningfully, e.g. to locate the PD source or identify the mechanism involved. PSPDA shows that the shape of the PD pulses supplies additional information regarding the transmission path of the PD pulses [3]. The PD shape information can hence help to increase the localization of the PD pulses.

An accurate model of the PD in transformers, motors and generators would enable better diagnostic results. The propagation of PD pulses through a generator would however require very advanced EM modelling capabilities [4]. Due to the physical problems and constraints involved, the initial modelling of smaller motors and transformers is proposed.

1.2 PROBLEM STATEMENT

The benefits of measurements relating to the condition of the insulation material, including PD measurements, are well known [9]. The calibration problem however restricts the general applicability of research. Various factors (such as insulation material, age, temperature, load and construction) are known to influence the recorded PD measurements. Sometimes more than one failure mechanism occur simultaneously, complicating the diagnosis. The electromagnetic influence of components in the generator is also currently unknown over the frequency range of typical PD pulses.

The correct interpretation of PD measurements is dependent on expert knowledge of the specific machinery [7]. The shortage of skilled personnel have prompted research into the automated on site processing and diagnosis of PD measurements [6]. While good results have been achieved under

laboratory conditions, no commercially available system can offer true diagnostic capability for practical systems.

Due to economic constraints, the generator data sets are relatively small in size and the PD status is sampled maybe once a year, depending on the severity of the problem. While it may be possible to diagnose the presence of a specific failure mechanism by trending the PD measurements, it is quite difficult to predict the actual condition, the source of the PD and to estimate the residual life of the machinery [9]. In order to solve the problem of statistical variability in sparse data sets and to enable better interpretation of the available data, it is proposed that a combination of experimental procedures, data based modelling, expert knowledge and automated classification be used.

A key part in this endeavour would be the systematic collection of a PD database, including PD shape analysis. Contin has illustrated the benefits of analysing the detailed PD waveform instead of just the accumulated charge in the modelling of PD phenomena [1,2].

The automatic classification of PD measurements can be implemented with a variety of techniques. Traditionally two types of technique were applied to problems of this nature:

- Rule based techniques are easy to understand. Translation of the knowledge of a domain expert into a rule set or expert system, sufficient to cover all cases, can be problematic. The rule set is frequently inflexible, as the number of rules grows exponentially with the complexity of the problem.
- Data based techniques use training data to derive a model explaining the input-output behaviour of the system. Large data sets are required to cover the multivariate feature space adequately, especially with non-linear relationships.

An expert system, while good at modelling logical structure and relationships, is not flexible in adapting to new data. Furthermore, it is difficult to extract accurate domain knowledge, and changes in the rule set can be problematic. The modelling of environmental conditions often requires many rules.

A PD database would require many measurements to adequately cover the multidimensional landscape. Alternatively a good model may be used to supplement the data set. The incorporation of a priori knowledge helps to reduce the data requirements, but the classifier also needs to adapt according to data. Some techniques can incorporate rules readily, but it is not so easy to incorporate quantitative models. By way of summary the following research objectives can be stated:

- The quantification of the factors (some directly observable and some not) influencing PD measurements
- The accurate modelling of PD signal generation and propagation in complex machinery such as large generators.
- Quantification of the rate of degradation in order to minimize potential loss of production.

2 DATA PROCESSING

2.1 Data collection

The collection of a consistent and reliable set of measurements, covering the physical and electrical properties of the machinery, is very important for a reliable diagnosis of faults [11]. A priori knowledge regarding the factors influencing PD measurements is important in order to define a meaningful and compact feature set. Statistical considerations indicate that the variance of results can be reduced by increasing the sample size. The sample size is especially important to discover the potential non-linear relationships inherent in the data. The discovery of variables with predictive power, using non-linear analysis, helps in the modelling process.

2.2 Linear and non-linear processing techniques

The benefits of non-linear analysis techniques versus standard linear data analysis techniques depend on the complexity of the problem. Neural network classifiers, with non-linear feature mapping using sigmoidal basis functions, can identify multiple failure mechanisms and noise in PD measurements [5]. The degree of non-linear correlation inherent in a data set is very important in the design of a classifier. It can be estimated by an entropy based algorithm such as the mutual information criterium [12]. This technique has been demonstrated in analysing the non-linear contribution of variables in a large plant.

If small data sets are available but many features are used overfitting of the model is highly likely. Regularization techniques can be employed to penalize the degree of non-linearity in the initial modelling stages. After progressing through a number of iterative data modelling and experimental verification stages, in which the confidence in the modelling process increases, the complexity and degree of non-linear interactions of the classifier can be gradually increased. The aim is to develop an accurate model for the automatic classifier, while keeping the model as simple as possible.

2.3 Feature selection and extraction

The PD measurements form a multivariate feature space. The required number of training samples for a classifier generally increases exponentially as function of the number of features, assuming uncorrelated data [5]. Furthermore, the performance of the classifier is closely linked to the quality of the features.

The extraction of a compact feature set, which can still capture most of the correlation inherent in the original sample space, is thus very important in a multivariate setting. Suitable feature extraction methods highlight the important discriminating characteristics of the data, while simultaneously ignoring the irrelevant attributes (i.e. noise) [5].

2.3.1 Trending of the data

Special techniques have been developed to establish long term trends in the data [12]. The selection of informative features according to PCA or information theoretic considerations can be misleading in the extraction of long term trends in the data. Short term behaviour, while being informative about the relations between different parameters, constitute noise in the determination of long term trends. The feature set should thus be designed to include features which may be very small in terms of variance, compared to the features capturing the short term behaviour, but which is aimed at capturing long term behaviour. The percentage variance contributed by the long term behaviour would generally constitute a very small percentage of the total variance. The need for specialised techniques is therefore apparent.

2.3.2 Weibull function

Experimental results have shown that if the insulation system is in steady state and only one mechanism is active, the two-parameter Weibull function

$$F(x; \alpha, \beta) = 1 - \exp \left[- \left(\frac{x}{\alpha} \right)^\beta \right]$$

provides a good approximation to the detected PDHD [1]. In the case of multiple discharge mechanisms, an additive Weibull model can be used to model the data:

$$F(x; \alpha_i, \beta_i) = \sum_{i=1}^k p_i \left[1 - \exp \left[- \left(\frac{x}{\alpha_i} \right)^{\beta_i} \right] \right]$$

2.3.3 Wavelet transform

Pre-processing by the wavelet transform has been shown to lead to increased compression ratios, better noise reduction and increased classification performance. The wavelet transform could hence be used to reduce the dimensionality of the pulse phase PD measurements.

3 PATTERN RECOGNITION

The performance of a non-linear classifier, such as a neural network, is closely dependent on the number of training examples relative to the degrees of freedom (complexity) of the classifier [5]. As a rule of thumb, the number of samples should be 10 to 100 for each independent feature. The combination of fuzzy logic based techniques with neural network techniques can lead to an improved classifier with the following advantages [13]:

- Fuzzy rule generation and inference provides more transparent understanding of the internal operation of the classifier, compared with purely neural classifiers. A priori expert knowledge of the PD mechanisms can be incorporated in the design phase as a set of fuzzy rules.
- Vagueness and ambiguities can be handled gracefully. The degree of membership with regard

to the output class provides an indication of the certainty of presence a mechanism, rather than a crisp 'present' or 'absent' answer.

- Multiple rule sets can be stored, e.g. different machines could use different rule sets.
- The output signal represents the degree of membership with regard to a specific class. By monitoring the confidence of the FLNN in a prediction, the extrapolation of the FLNN beyond training data can be established. In regions of the input space with sparse data the classifier could request additional training samples.
- Hardware implementation for fast online decision-making is available.
- The number of free parameters that can be adjusted is generally smaller than for the FFNN, reducing the chance of overfitting small data sets with non-linear relationships and reducing the requirement for training samples.

3.1 Experimental design

The reduction of the feature set to the minimum required for acceptable modelling is important in the design of structured experiments. The influence of a set of experimental variables on the response variable(s) is determined by conducting a series of experiments. The resulting response surface can be used in a simulation to augment the existing data set. Since the required number of experiments is exponentially proportional to the number of experimental variables to be considered, the smallest possible number of features should be used. An exhaustive evaluation of all possible experiments would be prohibitive. More sophisticated experimental design techniques would therefore be required in order to minimize the required number of hypercubes to cover the multivariate space.

In practice the generator is a critical part of the plant and it would be very difficult and costly to interfere with the normal mode of operation, precluding the steering of the plant in order to obtain measurements.

4 CONDITION MONITORING

4.1 Analysis of data set from plant

A set of 130 PD measurements for 6 electrical machines, with 2D pulse phase distributions, has been obtained. Each machine represents a different failure mechanism. Machine D had 56 PD measurements for the 3 phases, or 18 PD measurements per phase. Only 3 measurements are available before rewinding, which is insufficient even to check the diagnosis of the rule based method. After rewinding, the 9 available measurements consisted mostly of system tests and calibration. A month later another set of 7 measurements, at two different temperatures, became available. It is clear that the measurements, although not invalidating the rule based approach in any way, is insufficient to enable accurate modelling by any data based approach.

To overcome this problem a model can be developed to expand the set of measurements. Due to the difficulty in validating a model for PD measurements in large generators, it is proposed that a simpler model for a smaller motor or transformer be generated first. After verification and possible improvements, additional data can be generated from the model and tested against the actual transformer. Since such an experiment can be controlled and manipulated, verification is simplified considerably. The experiment could then be repeated with other similar machinery to build up a PD database. The performance of various techniques could also be assessed more objectively than in the case of a turbo-generator.

4.2 Calibration of PD measurements

The calibration problem [7] remains one of the major obstacles in the universal application of PD measurements. Good results are reported for classification on specific machinery, but the general applicability of the results is not yet known. The building of a PD database depends on the accumulation of PD signals from various machines, operating conditions, manufacturing standards, measuring techniques, etc. The influence of these parameters on classification performance is not clear. A set of rules based on certain PD parameters is available and different failure mechanisms can be identified. Unfortunately, no quantitative indication of the current machine condition is provided.

4.3 Shape of PD pulses

Initial PD measurements used the PDHD. The PD pulse phase analysis added another dimension, which made recognition of failure mechanisms easier. Since the PD pulses are much shorter than the noise pulses, it is possible to filter the noise digitally. More recently, the waveform of the PD pulses has been shown to further improve the diagnostic performance [3].

4.4 Sampling of PD pulses

Recording the waveform of the PD pulses is very memory intensive due to the high sampling rates involved (around 1GHz). It is therefore appropriate to consider more advanced sampling techniques in order to minimize the storage requirement without impairing performance [2]. A PSPDA (pulse-shape partial discharge analysis) system was developed to analyse the observed PD pulses. The system can provide information relating to the shape characteristics and thus lead to improved PD pattern analysis [2].

4.5 Localization of PD pulses - development of a high frequency EM model

The localization of PD pulses in electrodes, rotating machines and cables is described in [10]. A PD pulse causes a travelling wave, which is reflected at the end of the cable, returns and is then again reflected at the other end. A series of impulses of decreasing height is formed and can be measured by a detector with sufficiently small time constant. The location is

proportional to the time difference. An accurate high frequency model of the generator would allow the prediction of the waveform of a PD pulse after propagating through the system. Unfortunately, the lumped capacitor model employed in transmission lines is not valid for frequencies above 450 kHz [8]. The development of a high frequency model would be very desirable, but also quite complex. The influence of several factors are still unknown, such as the effect of mutual inductances and capacitances, the role of the insulation system, impedance mismatching and the dynamics which could lead to different transfer functions, depending on the location [4].

5 CONCLUSION

Various aspects relating to a successful condition-monitoring program based on PD measurements have been described. The construction of a quality PD database would enable accurate non-linear relationships to be identified. The development of an accurate high frequency EM model would enable accurate localization of the PD pulses.

An automatic classifier for PD recognition should be a combination of traditional rule based and data based techniques, combining the strengths of both techniques. In this way overall system performance can be optimised by incorporating expert knowledge as well as available measurements in the design phase.

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ION CLOUD COUPLING BETWEEN HIGH VOLTAGE DIRECT CURRENT TRANSMISSION LINES AND ADJACENT CONDUCTORS

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Abstract: Eskom have previously experienced an undesired ionic coupling effect between telephone lines and the Cahora Bassa high voltage direct current transmission lines. This project was initiated to provide Eskom and Telkom with information on the mechanism, impact and influencing factors to bring them closer to a possible solution. It consists of a through literature study on the subject of ions and its source, corona concerning HVDC transmission lines. High voltage laboratory work was done on a model to investigate the relationship between the ion current and the electrostatically induced voltage. Emphasis was put on the influence of relative humidity, solar radiation and wind direction. It also deals with the levels of ion current as a function of the applied voltage.

Keywords: *Corona* is a luminous discharge due to ionisation of the air surrounding an electrode and occurs when the applied voltage exceeds a certain critical value.

Ions are formed when electrons tears loose from a neutral gas molecules leaving behind a positively charged molecule, which is called an positive ion.

1. INTRODUCTION

Since the re-commissioning of the Cahora Bassa $\pm 533\text{kV}$ HVDC Transmission System in 1998 previously encountered operational problems are been assessed by Eskom. One of these problems is the corona-enhanced ionic coupling to open-wire telephone lines. Clouds of space charge are caused by corona on the conductors. These clouds of ions then literally drift away, some of it will collect on telephone lines, thereby inducing thousands of volts which then become dangerous to Telkom linesmen and exchange maintenance staff. This project will provide Eskom and Telkom with information on the mechanism and impact of the phenomenon to develop safe working procedures.

2. HVDC TRANSMISSION

Hydroelectricity has become an ideal options for electricity generation because of its clean emissions and a reduced environmental impact. Unfortunately economical locations are very limited since most of the viable locations not already utilised is in Africa with no nearby loads. However, using HVDC transmission, energy can be transported over long distances at much lower cost than HVAC thereby transmission to say Western Europe could become economically feasible.

3. IONIZATION

A free electron between two electrodes will be accelerated towards the anode by the electric field. If such a electron attains enough speed it will transfer sufficient kinetic energy during a collision with a gas molecule to cause ionisation. One or more electrons could leave the molecules orbit to also be accelerated

toward the anode partaking in further collisions, thus the number of electrons and positive ions increase exponentially and it's called an avalanche.

3.1 Self-sustaining discharges: Townsend discharge

A self-sustaining discharge occurs when there is a source of electrons to form repetitive avalanches. In accordance with the Townsend mechanism those electrons are supplied by the positive ions when they collide with the cathode. If one electron starts at the cathode it will form an avalanche with the number of electrons released.

$$n = e^{\alpha x} \quad \dots(1)$$

When these electrons have been absorbed by the anode, the positive ions with their low mobility have hardly gained any speed. When they collide with the cathode, each one have a certain probability of releasing electrons depending on the kinetic energy gained on it's way to the cathode. The number of electrons released by the impact of one positive ion with the cathode is represented by Townsend's second ionisation coefficient (γ). These new electrons emitted will form an growing number of avalanches.

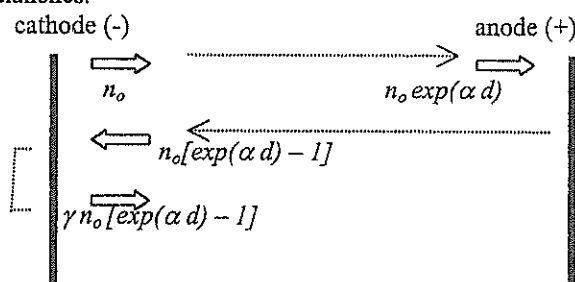


Figure 1 Secondary ionisation

Considering figure 1 the number of resulting ions due to collisions are

$$n_o \exp(\alpha d) - n_o = n_o [\exp(\alpha d) - 1] \quad \dots(2)$$

Therefore the number of electrons released during impact with the cathode are given by

$$\gamma n_o [\exp(\alpha d) - 1] \quad \dots(3)$$

For this process to be self-sustaining there should be a gain of free electrons, i.e. $\gamma n_o [\exp(\alpha d) - 1] > 0$

3.2 Ion Currents

Ion current density is usually detected by a Wilson Plate (usually 1m^2) lying in or slightly above the ground plane. The plate is insulated from ground and the current is measured with an electrometer type circuit which place the probe plate at a virtual ground reference.

Near HVDC lines ion currents has three components:

i. Displacement current

The electrostatic field under a HVDC line changes due to the space charge as ions drift in the wind. This will produce a random displacement current.

ii. Conduction current

The conductive current is caused by the motion of ions in the electric field and is directly related to the electric field (E) and charge density (ρ) by $J = \rho k E$ where k is the ion mobility.

iii. Convection current

The convection current is due to the transport of ions and charged aerosols by the wind.

Since HVDC transmission lines generally operate continuously above the corona onset voltage, considerable space charge is generated. This ions will flow from the conductors to the ground plane or other conducting bodies, thereby inducing a voltage which is directly proportional to the body's resistance to ground.

4. CORONA

Corona is the source of ion currents. It occurs due to the high concentration of electric field lines near electrodes of small radius of curvature. Corona is limited to these areas and is a self-sustaining discharge.

Electron avalanches between electrodes will cause positive and negative ions to collect in those areas. These ion clouds are referred to as space charge. For AC the charge will be removed by the opposite polarity every half cycle. DC corona will cause space charge to fill the area between electrodes, and from electrodes to

ground. There is a distinct difference between corona around negative or positive electrodes.

4.1 Negative Corona

Trichel-pulses occur if an applied voltage is raised above the corona inception gradient. In air, the pulses are regular recurrent and constant in size. The regular character of the pulses can be explained in the following way. The process starts when a Townsend discharge takes place at the cathode. This avalanche will create photons which will generate even more free electrons by photo-ionization. Further from the cathode, the electrons will attach themselves to the electro-negative oxygen molecules in the air. The ionized region therefore quickly expands, and shields the point of corona from the electrical field. This will extinguish the discharge for 50 to 200 μs . During this period the electric field will move the negative charge away, causing the field to rise again and the cycle is repeated.

4.2 Positive Corona

Streamers are formed near the point of corona. These streamers leave positive ions behind, which will shield the point and therefore extinguish the discharge. This space charge will then drift away and a recurrent discharge pattern is formed. The streamers do not have the same regular pattern as negative corona and they will increase in length as the applied voltage increases.

4.3 Corona sources on HVDC transmission lines

Transmission lines are subjected to diverse atmospheric conditions, which change all the time. Some of these changes have a very definite effect on the corona performance of DC lines, while others seem to have very little effect. Dust, raindrops and snow on the conductors will change the electric field and will all become sources of corona. Other corona sources on conductors are nicks, scratches, gouges and dead insects. EPRI found that dead insects were a major source of corona [1]. It was found in large numbers on the positive conductor all year except for the winter.

Line parameters also influence corona losses. Bipolar losses are estimated to be 3 to 5 times the monopolar losses on a circuit basis [2]. Corona losses decrease as the pole spacing between conductors of opposite polarity increases. An increase in line height will also decrease the corona loss.

DC lines will exhibit a much smaller change in corona loss when compared to AC lines in the transition from fair to rain weather. Typically, DC losses may increase

by a ratio of 10 to 1 while AC losses may increase by 50 to 1.

5. CORONA, IONS AND THE ELECTRIC FIELD

The effective electric field of a HVDC transmission line is determined by the voltage on the transmission line, the charge (ions) generated by corona on the conductors and charged aerosols. The corona-generated ions have the same polarity as the conductor and will drift away from the conductor to either ground or a conductor of opposite polarity. Space charge of the same polarity under a conductor will increase the electric field at ground level, while that of opposite polarity decreases the electric field. These corona generated space charge make the electric field dynamic, complex and difficult to predict.

The electric field due to the voltage on the conductors is known as the electrostatic field and follows the variations of the line voltage, therefore except for a small ripple, it is constant in time and direction. The electric field due to the charge released into space (space charge) is known as the space charge field. Because corona is variable in nature, and wind randomly move charges in space, the space charge field is fluctuating. The superposition of the electric and electrostatic fields produce the effective electric field. The electric environment in proximity of a HVDC transmission line depends on many parameters such as: voltage, conductor height, pole spacing, conductor dimensions, conductor surface conditions, weather conditions, season, climate, aerosol content of air, wind speed and direction, type of terrain (i.e. its geometry and corona emitting characteristics).

6. LABORATORY TESTS

6.1 Voltage Source

A half wave HVDC construction set was used for the tests. The peak to peak ripple voltage is reduced to be less than 2% of the total voltage, which is necessary to reduce the AC coupling effect.

6.2 Scale Model

The model consisted of two copper rods suspended from a PVC framework by thin nylon lines to maximise the resistance to ground. Conducting spheres were fitted to the ends of the copper rods to reduce the extremely high field at these sharp ends causing discharge to the wall. The rods was 1000 mm in length and 250 mm apart. HVDC was applied to the one rod which was 400 mm from the ground plane while the other rod, which collected the ions, was 200 mm from the ground plane.

7. MEASUREMENTS OBTAINED AND DISCUSSION

7.1 Effect of Wind

The space charge flows from conductors to ground or between conductors. This ion movement is restricted to the electric field lines in still air conditions. The ions initially have high velocities that will decrease rapidly with the electric field due to its distance from the conductor. At ground level and interelectrode areas the velocities are typically in the region of 3m/s [2]. Wind velocities are typically in the same range and tend to move ions from the lines of force, downwind.

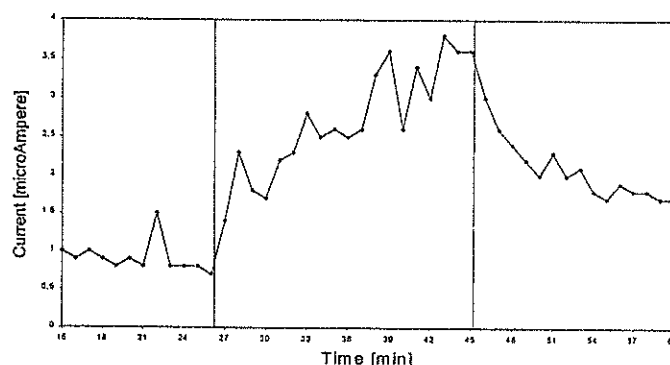
Ion clouds create a stable atmosphere around the conductors since the polarity of ions suppresses further formation of ions. The wind will free large numbers of ions from the constraints of the electric field, which will cause more ions to form, thereby restoring stable conditions. As the wind thus increases, corona currents and corona loss will increase. It was also observed that an increase in the applied voltage would decrease the influence of wind. This is since the increase in the electric field will increase the velocities of the ions making it less susceptible to the wind.

7.2 Effect of Humidity

Very little research has been done which investigated the influence of humidity on the ions. Steam was used to simulate this effect. Care was taken not to change the temperature of the air and to limit condensation on the conductors. The latter is necessary since water droplets would become corona sources.

The first test was done at a voltage of negative 70 kV and an initial humidity of 65%. After 26 minutes the humidity was increased until it reached 85% during the 45th minute. The steam source was then removed and the humidity started decreasing, reaching 75% at 60 minutes. The results are shown in figure 2.

Figure 2 Effect of Humidity at -70kV



It is clear that the humidity has an effect since the ion current increased proportional to it. This is a significant result since there has been some uncertainty whether humidity has an effect or not. Ionisation will increase since H₂O (12.6eV) requires less ionisation energy than N₂ (15.576eV). The ionic coupling will also increase since the water molecules are polarised, this will cause them to transfer charge from one conductor to the other.

7.3 Effect of Solar Radiation

It was observed that direct sunlight had an influence on the magnitude of the induction and its frequency of occurrence. The minimum amount of energy required to remove an electron from the ground state of an atom is called the ionisation energy. Ionisation occurs when

$$\lambda \leq \frac{ch}{W} \quad \dots(4)$$

Where, h = Planck's constant,
 c = the velocity of light,
 λ = wavelength of the incident atom, and
 W = ionisation energy of the atom.

Thus, the higher the ionisation energy the shorter wavelength of radiation must be for ionisation to take place. The ultraviolet spectrum is conventionally divided into three regions namely UV-A from 315 to 400 nm, UV-B from 280 to 315 nm and UV-C from 100 to 280 nm. No solar radiation below about 290 nm reaches the earth's surface [3].

It is expected that the ionisation and thereby the ionic coupled current would increase under direct sunlight. This is since solar radiation makes free electrons available, which are necessary to start an avalanche. However no difference could be obtained with a UV-B source. Using equation 4 it was then found that that UV-B can only supply ± 4 to 4.5eV, which is not high enough to ionise either of the two majority air molecules namely, O₂ (12.063eV) and N₂ (15.576eV). Therefore no real difference should be obtained and this highlights a very common error that solar radiation will cause ionisation in air.

8. CONCLUSIONS AND RECOMMENDATIONS

The relationship between the ion current and the electrostatically induced voltage depends on the impedance of the floating line. The higher resistance to ground of the open ended wire, the higher will induce voltage become, simply from ohms law. The capacitance of the line will have no great effect on the magnitude of the current and voltage. It merely effects

the time constants, i.e. the time to reach a minimum or maximum value

The wind velocity is an important factor, and as it increases, so will its effect. Humidity has a definite influence and will increase the ionic coupling. It is recommended that this test should be done inside a closed chamber using salt solutions to obtain different humidities at which the ion currents can be compared. An ion counter can be used to determine the increase in ionisation and the test could be done in a noble gas such as Ar to determine the severity of charge transfer. The results of the test on solar radiation did not prove an increase in the ion current. This is since the high energy UV-B wave do not possess enough energy to cause ionisation. It would therefore seem that direct sunlight has no effect due to its radiation but it might influence the ionisation current due to a temperature increase. The effect of temperature should be investigated to confirm this.

The ion current density was very dependent on the applied voltage. Below the corona inception voltage, no current was measured, but once above this gradient the current increased proportional to the voltage applied. It is recommended that for future research a datalogger should be used to establish a more average measurement result since it fluctuates a lot.

ACKNOWLEDGEMENTS

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PRELIMINARY RESULTS FROM AN EXPERIMENTAL INVESTIGATION INTO THE GROWTH OF WATER TREES UNDER 50Hz SIMULATION IN SAMPLES OF PRODUCTION CROSS-LINKED POLYETHYLENE CABLES

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Abstract: In this paper the results of the first phase of a study to grow water trees in XLPE cable insulation are given. Needle-shaped depressions are made in the insulation of an XLPE cable sample, which is then submerged in water while a voltage at power frequency (50 Hz), is applied. The results show that consistent water tree growth did not occur. Due to problems encountered while cutting the insulation in the lathe, the relative position of the depressions where the largest water trees occurred could not be determined.

Keywords: XLPE, Water trees.

1 INTRODUCTION

The failure of much of the first cross-linked polyethylene (XLPE) cable manufactured in South Africa has been attributed to water trees that grew from the loosely extruded core screen of the cable [1]. In another investigation it is reported that all the samples of service aged cables removed and examined contained electrochemical (vented) trees. The large trees were all growing from protrusions on the extruded core screen. The protrusions were all sharp, resulting in high stress enhancement at each tree site [2]. Where the breakdown of the cable insulation could be attributed to trees, it was also found that electrical trees appeared together with water trees [3].

The main causes of electrical aging in cross-linked polyethylene insulated cables are water trees, electrical trees and partial discharges [3]. The voltages required to degrade cross-linked polyethylene insulation are very much lower than the breakdown strength of the insulation, and the degradation due to these processes is therefore termed low level degradation. As partial discharges occur only in microscopic voids, it is the water trees and electrical trees that lead to the most damaging deterioration in the insulation of good quality cables.

Due to the fact that the growth of water trees under normal service conditions is a very slow process, it is necessary to accelerate their growth for experiments to be performed in laboratories. Most of the procedures described [3] are performed on samples of insulation and not on the insulation of a production cable. In this study, lengths of production cable, submerged in water are used for tests in the laboratory.

2 NEEDLE-SHAPED DEPRESSIONS

In order to accelerate the process of tree growth, needle electrodes are employed to increase the field from kV/mm to nearly MV/mm [3]. For accelerated growth of electrical trees, double needles or point-plane geometries are used. In the case of accelerated growth of water trees, needle shaped depressions filled with water are used.

Figure 1 shows the dimensions of the sample of the production cable that was used in these experiments and Figure 2 a scanned image of the needle-shaped depression in the XLPE insulation through the extruded semi-conductive screen.

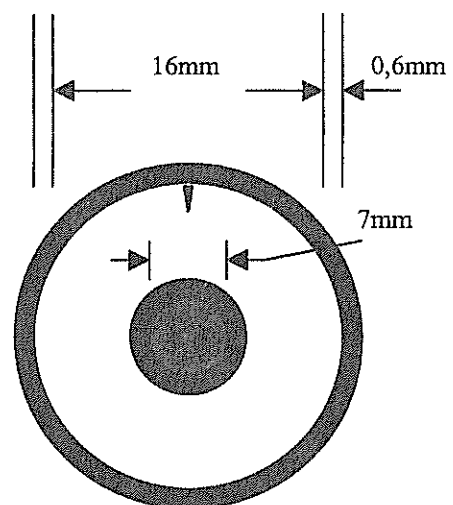


Figure 1. Dimensions of cable sample

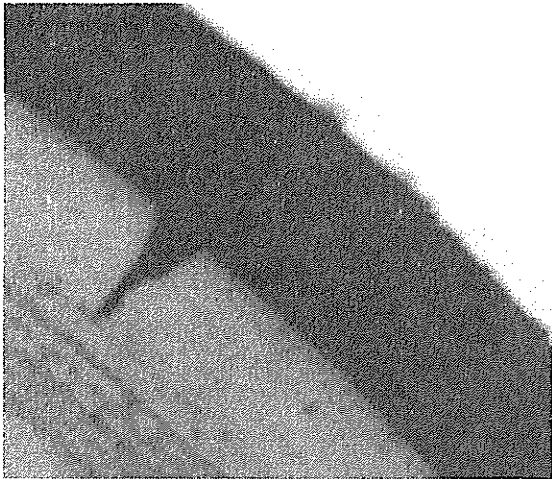


Figure 2. Scanned image of needle-shape depression.

3 PREPARATION OF SAMPLE

The outer PVC sheath, armouring and copper tape are removed to expose the outer extruded semi-conducting screen. The needle-shaped depressions are made through the semi-conducting screen into the XLPE insulation, allowing water to penetrate into the insulation when the sample is submerged.

4 TEST SET-UP

As this is the first phase of a series of experiments where the water trees grown in the XLPE insulation are to be subjected to voltage surges, it was decided to grow water trees in two locations on the cable and to compare the growth in the two locations. Ten needle-shaped depressions were made at each location (Figure 3), and then submerged in water in two separate containers as shown in Figure 4.

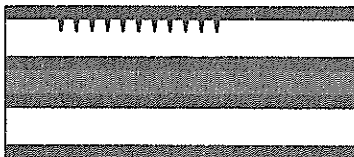


Figure 3. Ten depressions in sample

Not shown in the sketch are the re-usable cable terminations fitted to each end of the cable.

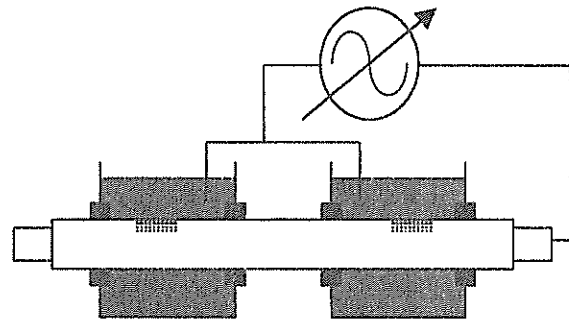


Figure 4. Test set-up

Depth of needle-shape depression: 0,45mm

Distance between depressions: 2mm

Applied voltage: 20kV

Duration of test: 8 days (192 hrs)

The temperature of the water was not controlled and varied according to the changes in the ambient temperature.

Maximum stress at the conductor screen:

$$E_{\max} = \frac{20}{3,5 \ln \frac{8}{3,5}} = 6,91 \text{ kV/mm}$$

Stress close to core screen, before creating needle shaped depression:

$$E_x = \frac{20}{7,55 \ln \frac{8}{3,5}} = 3,2 \text{ kV/mm}$$

The stress at the tip of the needle-shaped depression will be much higher than the calculated values above, and will depend on the radius of curvature of the depression.

After completion of the test, the XLPE insulation where the impressions were made, was cut in a lathe into a spiral with a thickness of 0,75mm. The pieces of the spiral with the impressions were then identified and cut out so that each impression was captured in a disc.

The next step was to boil the discs in a solution of 0,1g of Rhodamine dye in one litre of distilled water for 15 minutes [4].

The discs were then examined under a microscope for water trees and the images captured by scanning the discs in a high-resolution film scanner. As no magnified measuring scale was available, a printout of the magnified impression and water trees was made and measurements done on the printout. The lengths of the water trees could now be calculated by taking the thickness of the extruded semi-conductor screen as reference.

5 RESULTS

Number of depressions	: 20
Number with comparable trees	: 4
Number with only 'side' trees	: 3
Number with small trees	: 8
Number with no trees	: 5

Figure 5 shows the scanned image of one of the four depressions at which comparable water tree growth was found.

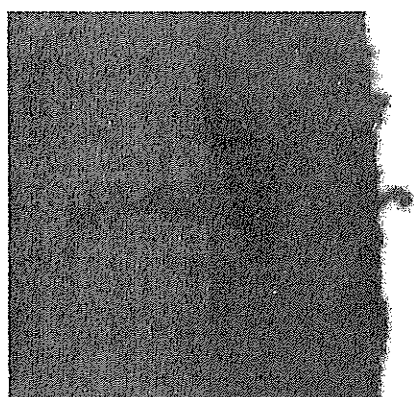


Figure 5. Depression with water trees

From the printout of the scanned image, the dimensions of the water trees in this sample are as follows:

Thickness of extruded screen	: 0,6mm
Length of tree at tip	: 94 μ m
Width of tree at tip	: 134 μ m
Length of tree at side	: 110 μ m

The dimensions of the water trees in the other three samples with the most growth are similar to the above.

A depression with a water tree growing from the side (a 'side' tree), is shown in Figure 6. When examining the samples with 'side' trees under a microscope, no sharp edges or points could be found at the base of the trees.



Figure 6. Depression with 'side' tree

6 CONCLUSION

These preliminary results show that water trees can be grown from needle-shaped depressions in a laboratory environment. The inconsistent growth in adjacent depressions can not be explained as yet, although one line of thought is that with the depressions very close, the depression that is slightly deeper shields the adjacent sites.

Due to certain problems encountered, the discs could not be marked and the relative position of each depression could therefore not be noted. A specific pattern for the water trees at the different locations could therefore not be determined.

The next set of experiments will be done using the same test set-up as shown in Figure 4. A maximum of five (5) depressions will be made in each sample and the spacing between the depressions increased to 30mm to reduce the likelihood of a depression shielding the adjacent sites. The applied voltage and time duration of the test will be the same.

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THE POLLUTION FLASHOVER OF AC ENERGISED CAP AND PIN INSULATORS: THE ROLE OF SHORTENING OF THE ARC.

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Consider the energised insulator, shown in Figure 1.

Abstract: This paper investigates the application of models to predict the flashover of practical cap and pin insulators. It has been previously shown that both DC and AC models can be used to predict flashover. The models are applied to the published flashover data of a number of different insulator designs and effective arc constants are obtained. It is found that the arc constants vary with pollution severity and that this variation can be attributed to the arc across the dry bands not following the insulator surface, but taking shorter routes.

Keywords: Insulator Pollution.

Introduction:

Pollution flashover models are usually based on simplifying assumptions such as a regular cylindrical shape or a uniformly distributed resistive layer on the insulator surface. These models are often modified, using the form factor concept, to allow for insulators with intricate shapes. In these models it is implied that the arc follows the contours of the insulator surface. In practice, however, insulators have intricate shapes and inter-disc or air breakdown may occur [1].

Lightly polluted insulators require a high stress to initiate pollution flashover. It is therefore conceivable that, prior to flashover, inter-disc or inter-rib air breakdowns may occur, thus shortening the arc length.

On the other hand, heavily polluted insulators require a low stress to flash over. The associated current levels are however high and, depending on the insulator design, the arc may be pushed out radially by the thermodynamic effect or inwards due to the electrodynamic effect [2].

Flashover Models for Insulators with arbitrary shapes.

The flashover processes of polluted insulators have been dealt with exhaustively in the literature [3], but the aspects, relating to arc shortening are highlighted in this section.

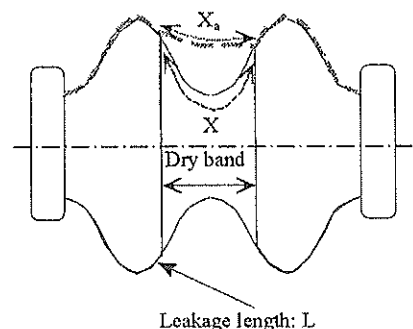


Figure 1: Leakage current flow and arcing over the surface of an insulator of arbitrary shape

The insulator surface is covered by a pollution layer that is conducting, except in the area of the dry band. As the electric stress can be excessive across the dry band, arcing occurs. If the length of the leakage path is L and the width of the dry band is X , the length of the conducting path will be $L-X$. At the points where the arc terminates on the pollution layer, the arc roots, the current flow will be nonuniform. If these current concentrations at the arc roots are ignored, the insulator, shown in Figure 1, can be converted to an equivalent cylindrical insulator, shown in Figure 2, using the form factor concept.

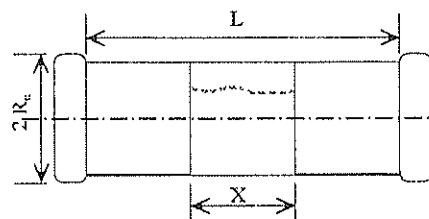


Figure 2 : Equivalent cylindrical insulator to represent the insulator in Figure 1.

Thus, the radius for the equivalent cylinder, representing the original insulator is given by

$$R_c = \frac{L}{2\pi F} \quad (1)$$

with F the form factor and L the leakage length, measured along the surface of the insulator of Fig. 1. Also

$$F = \int_0^L \frac{1}{2\pi r(s)} ds \quad (2)$$

with $r(s)$ the radius at position s , along the surface of the insulator.

The equivalent cylinder model is considered applicable, also in the case of a string of cap and pin insulators as, just before flashover the arc bridges a large portion of the string. The energised insulator is therefore represented by the equivalent circuit shown in Fig. 3.

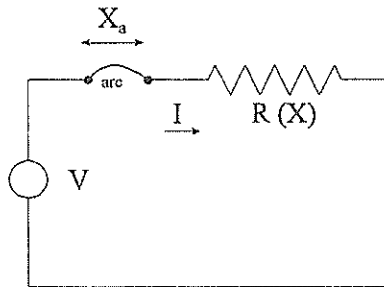


Figure 3 : Equivalent circuit for the energised cylindrical insulator.

The arc across the dry band is in series with the linear resistance, $R(X)$, representing the conducting layer of electrolyte. The Kirchhoff voltage equation for the circuit is

$$V = V_a + R(X)I \quad (3)$$

The arc has a nonlinear, negative resistance type of characteristic that can be represented by the following equation:

$$V_a = AX_a I^{-a} \quad (4)$$

where V_a : arc voltage (in V)
 A, a : arc constants
 X_a : arc length (in m)
 I : arc current (in V)

The arc length, X_a , is not necessarily equal to the dry band length, X . The relation between X_a and X is thus given by

$$X_a = k_a X \quad (5)$$

where k_a is the arc length reduction factor, such that $0 < k_a < 1$. This factor allows for arc bridging between sheds or discs on practical insulators. Under certain conditions, such as arc buoyancy, it is possible for k_a to exceed unity.

The resistance $R(X)$ depends on the following factors:

- The shape of the insulator, as represented by the form factor
- The pollution and dampness distribution on the insulator surface
- The current concentration at the arc root.

In order to obtain a simple solution, it is assumed that the resistance $R(X)$ is a linear function of X , i.e.

$$R(X) = \frac{(L-X)}{2\pi R_c \sigma_s} = \frac{F(L-X)}{L \sigma_s} \quad (6)$$

where: F : form factor
 σ_s : surface conductivity

Combining equations (1) to (6):

$$V = Ak_a XI^{-a} + \frac{F(L-X)I}{L \sigma_s} \quad (7)$$

An analysis of the interrelationship of V , I , X and σ_s shows that I and X increase without bounds if the following critical condition is reached:

$$\sigma_{sc} = F \left(\frac{S_c}{817} \right)^{\frac{a+1}{a}} \frac{(k_a A)^{\frac{1}{a}}}{L} \quad (8)$$

In this equation S_c in mm/ kV is the specific creepage length, given by the following equation:

$$S_c = \frac{1000\sqrt{2}L}{\sqrt{3}V} = \frac{817}{V} \quad (9)$$

The theory underlying this equation presupposes direct current and the approximations outlined above. When applied to AC, however, the voltage, V , represents the peak voltage. In this case, the value of the arc constant, A , allows for the incomplete thermalization of the arc. It also allows for the practical factors, mentioned earlier.

AC reignition model

A more accurate approach for AC energised insulators is to include the reignition characteristics of the arc into the above model. In addition to eq. (7), the following reignition constraint must also be taken into account, i.e.

$$V = NX_a I^{-a} \quad (10)$$

In this case the following solution follows:

$$\sigma_{sc} = \frac{Fn}{Lk_a(N-A)} \left(\frac{k_a NS_c}{(n+1)817} \right)^{\frac{n+1}{n}} \quad (11)$$

Eqs. (8) and (11) have similar forms and, as far as the flashover stress is concerned, are equivalent to each other, provided the correct numerical values are used for the constants [3].

It is interesting to note that, in both these equations, the factor k_a always appears next to either A or N . The effect of arc shortening

therefore results in effective values of A and N , i.e. $A_{eff} = k_a A$ and $N_{eff} = k_a N$.

Comparison of Model with Measured Flashover Data

There do not exist many references that give reliable test data of pollution tests, performed over a range of pollution and stress ranges. The test data of a number of insulators have been published by EPRI [4]. Diagrams of some of the tested cap and pin insulators are shown in Fig. 4.

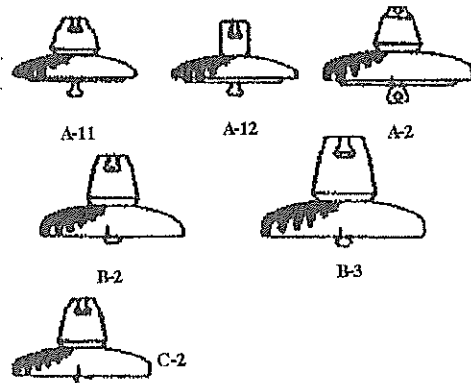


Figure 4 : Diagrams of tested insulators[4]

The flashover data of these insulators are given in Fig. 5. In this figure the pollution levels, given in the original publication in SDD (mg/cm^2) are converted to surface conductivity, using the conversion: $10 \mu\text{S} = 0.1 \text{ mg}/\text{cm}^2$ [5]. The 50 % flashover voltage is given in terms of the specific creepage length. The insulators with the higher graphs have the worst pollution performance.

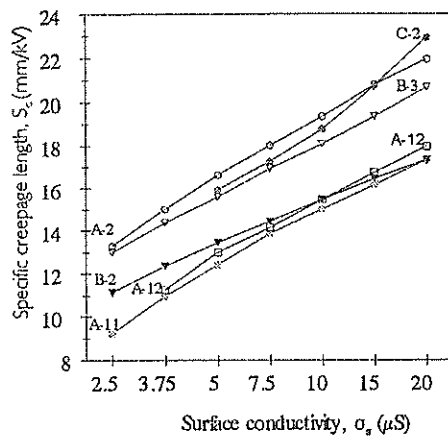


Figure 5 :Flashover data for the insulators of Fig. 4.

Eqs. (8) and (11) are now used to find the effective values of A and N , corresponding to the data in Fig. 5. The solution of eq. (11) is given in Fig. 6 for the AC model. In this case the following fixed values were assumed: $A =$

10, $n = a = 0.5$. The results for the DC model, i.e. eq. (8) are given in Fig. 7.

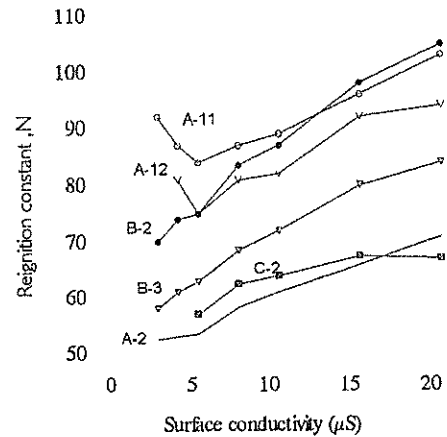


Figure 6 : AC model: Reignition constants for the various insulators versus pollution

Discussion of Results:

- Clearly, the graphs of Figures 6 and 7 are similar, confirming the equivalence of the DC and AC models as tools to estimate insulator pollution flashover.
- The values of N (in Fig. 6) or A (in Fig. 7) are functions of the surface conductance. Typically, a sharp reduction of A with decreasing surface conductivity in Fig. 7, suggests a low value of k_a , i.e. arc shortening.
- Arc shortening may be caused by inter-disc or inter-rib air breakdown. This is more likely to occur in the case of low surface conductivity, i.e. light pollution.

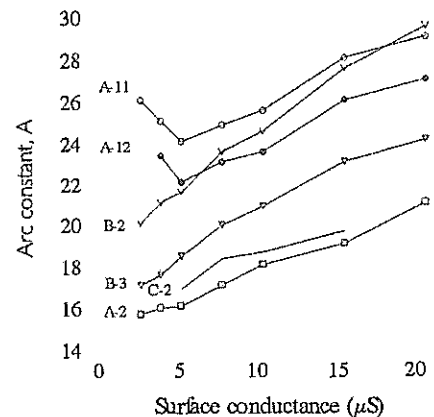


Figure 7 : DC model: Arc constants for the various insulators versus pollution level.

- The influence of arc shortening is illustrated in Fig. 8 where the data of Fig. 7 are normalised with respect to the largest value of A .
- The insulators A-11 and A-12 have

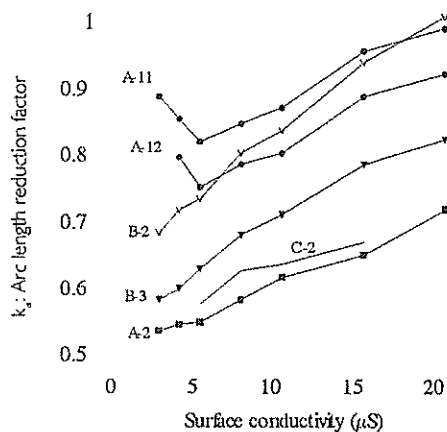


Figure 8: Arc length reduction factor, derived from Fig. 7

- the best pollution performance due to a lower degree of arc shortening, as is also confirmed in Figures 6 and 7. This is due to the low inter-rib factor
- The difference in performance of insulators B-2 and C-2 cannot be explained as the given data suggest that they are almost identical. However, a different inter-disc factor is given for B-2, that conflicts with the dimensions provided.

Conclusion:

It appears that the vastly different pollution flashover performance of cap and pin insulators may possibly be attributed to the occurrence of arc shortening during the flashover process. It is also clear that it is not feasible to use fixed values for the arc constants over the full range of stresses. The concept of effective arc constant has thus been broadened, also to include the stress-dependant arc shortening.

Acknowledgements:

Eskom, and TESP in particular, is thanked for support of this research.

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Table 1: Dimensions of the insulators in Fig. 4.

	A-11	A-12	A-2	B-2	B-3	C-2
Disc radius (R_M)	127	127	145	140	160	140
Axial spacing between discs (L_S)	146	130	178	165	198	172
Total leakage path (L)	305	305	395	370	425	370
Top leakage path (L_T)	109	109	124	124	131	120
Pin cavity radius (R_B)	20	20	20	20	20	20
Cap radius (R_M)	84	54	80	102	69	86
Bottom Leakage path (L_B)	196	196	271	246	294	250
Form factor (F)	0.76	0.81	0.94	0.88	0.91	0.88
Ratio: L/L_S : Inter-disc factor	2.08	2.34	2.2	2.32?	2.16	2.15
Ratio: $L_B / (R_M - R_B)$: Inter-rib factor	1.83	1.83	2.05	2.1	2.2	2.1

A PRELIMINARY CONSIDERATION OF THE EFFECTS OF MUTUAL INTERACTION BETWEEN COMPONENTS ON THE PERFORMANCE OF AN EARTH ELECTRODE SYSTEM SUBJECTED TO A LIGHTNING TRANSIENT CURRENT

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Abstract: A previously developed model of a site-wide earth electrode is used to further investigate the lightning transient performance of a typical earth electrode system that includes both concentrated and distributed elements. Non-linear effects of breakdown in the soil surrounding the earth electrode are included as well as the mutual interaction between components. A full-scale experimental arrangement is proposed in order to provide practical verification of the model predictions.

Keywords: Earth Electrode, Ionisation in Soil, Mutual Interaction of Conductors, Partial Lightning current

1. INTRODUCTION

Understanding the transient performance of an earth electrode is an important component of ensuring adequate lightning protection, especially for a site-wide earth electrode that forms part of the bonding and equipotentialisation network [1-3]. The earth electrode is used to provide a site-wide equipotential platform [4], but also provides a path for partial lightning current that can induce damaging overvoltages in signal and power cables in the vicinity [5, 6].

Initial modelling performed by Van Coller [7] was extended to produce a more detailed model that includes the effects of breakdown in the surrounding soil based on Liew's dynamic model [1, 8]. The focus of this paper is to investigate results obtained using the improved model for a site consisting of two interconnected buildings, with emphasis placed on obtaining the distribution and shape of partial lightning current in the system. The model is further improved by including the effects of mutual interaction between components of the earth electrode. Finally, an experimental arrangement is proposed in order to investigate the behaviour of the complete earth electrode system model. Therefore the paper takes the following format:

- Introduction
- Case study
- Effect of ionisation
- Mutual interaction
- Consideration of a suitable experimental procedure
- Discussion
- Conclusion

2. CASE STUDY

A large number of lightning related failures were occurring per year at the industrial site selected as the case study prior to the application of correct earthing, bonding and equipotentialisation. Hence the need to quantify and numerically investigate the performance of a large earth electrode system was identified [1].

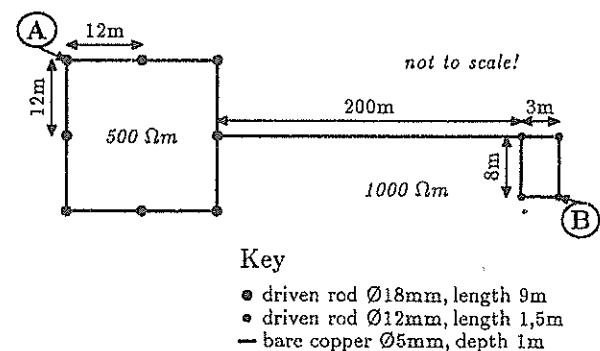


Figure 1 : Earth electrode system configuration for case study

The earth electrode system configuration of the industrial site is shown in Figure 1. The system has been modelled using equivalent circuit analysis and the MODELS extension to the widely accepted Alternative Transients Program (ATP) version of the Electromagnetic Transients Program (EMTP) [9]. The non-linear effects introduced by breakdown in the soil surrounding both concentrated and distributed elements of the earth electrode are included in the model and a lumped parameter transmission line model is used for distributed earth electrode components. Initially, the mutual effects be-

tween components was omitted. More specific details of the model implementation can be found elsewhere [1].

Throughout modelling, a standard 32 kA 8/20 μ s lightning impulse current waveform [10] was chosen to represent the injected lightning current. For simplicity, the effect of other components of the lightning protection system have been omitted (for example the down-conductors), as well as the possible interaction between the lightning channel and the earth electrode system [11].

Due to the relatively large length of the driven rod electrodes installed around building A (Figure 1), the self-inductance of the rods was included in the model. The resistivity of the soil was found to be 500 Ω m for building A and 1000 Ω m for the trench and building B using the Wenner method. As a simplification, the soil was assumed to be homogeneous and isotropic. The dielectric constant of the soil was assumed to be 4, corresponding to dry soil [12].

Of particular interest is the current division nearest the points of injection as shown in Figure 2. The current divides between the driven rod and the two branches of the trench earth. Point A and point B refer to the points in Figure 1. The current waveform along the trench is also important since it will determine the magnitude of induced voltages on neighbouring signal and power cables.

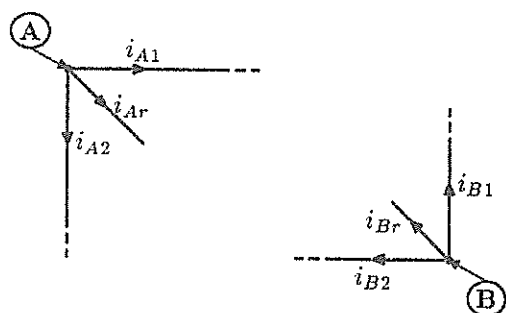


Figure 2 : Current division at selected injection points

3. EFFECT OF IONISATION

The non-linear effect of ionisation that occurs in the soil around the earth electrode (due to high current densities on the surface of the electrode) is commonly ignored as a simplification to modelling. This might be valid for very low soil resistivities, but for typical South African conditions this assumption is not valid [1]. To investigate the effect of this ionisation on partial lightning current distribution, a model using low current resistance values for all earth electrode elements was compared against a model that included the effects of ionisa-

tion - Figure 3.

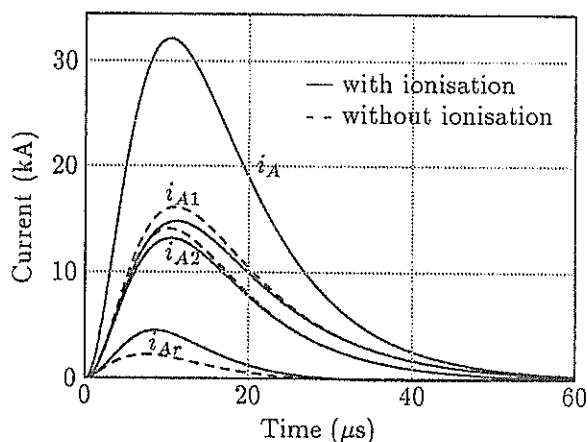


Figure 3 : The effect of breakdown in the soil surrounding the earth electrode system

Table 1 summarises the partial lightning current values of interest obtained from the modelling (the percentage of total injected current is also shown). The current entering the trench joining the two buildings is denoted $i_{trenchA}$ and the current entering building B at the end of the trench is denoted $i_{trenchB}$. Twice as much current is dissipated into the earth by the driven rod at the point of current injection due to ionisation effects. The difference in currents at the end of the trench earth are less marked due to the filtering effect of the distributed earth electrode.

Table 1 : Comparison between maximum current values for ionisation.

Current	With Ionisation		Without Ionisation	
i_A	32,4 kA	100%	32,4 kA	100%
i_{Ar}	4,6 kA	14%	2,3 kA	7%
i_{A1}	14,7 kA	45%	16,1 kA	50%
i_{A2}	13,1 kA	41%	14,0 kA	43%
$i_{trenchA}$	10,5 kA	32%	8,8 kA	27%
$i_{trenchB}$	1,4 kA	4,3%	1,2 kA	3,7%

4. MUTUAL INTERACTION

Previously ignored as a simplification to modelling [1,7], the effect of mutual coupling between components of an earth electrode system can be significant even for reasonably separated components. Two important observations can be made regarding the site shown in Figure 1. Firstly, there is no coupling between elements at right angles to each other. Secondly, all other elements that make up the site earth electrode are simply parallel conductor arrangements. Based on the assumption that the diameters of the various earth electrode components are small with respect to their length, the mutual

inductances can be calculated using the straight filament approximation [13,14]. The total inductance of each element is the sum of the self-inductance of the element and the mutual-inductances of any coupled elements.

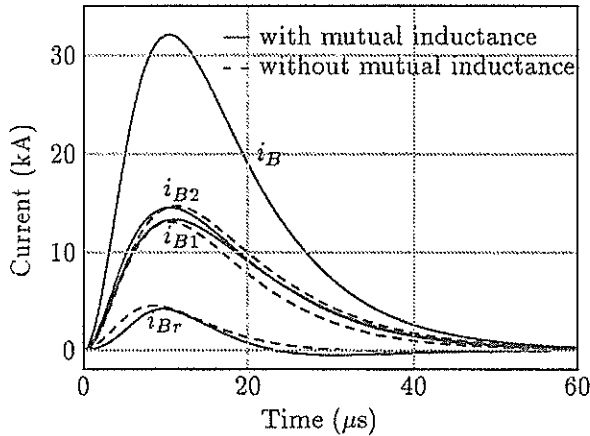


Figure 4 : Comparison between models including and excluding mutual inductance

Inclusion of the effects of mutual inductance in the improved model resulted in no significant difference for current injection at point A. This is because the earth electrode components of building A are well spaced. However, *Figure 4* shows the results for current injection at point B. The relatively small dimensions of the earth electrode system around building B result in the mutual coupling effects being more pronounced and the clear effect on the shape of the partial lightning current waveform can be seen.

It is interesting to note that when the point of injection is the driven rod at point B, a current of 4,2 kA is dispersed into the driven rod, which is comparable to the 4,6 kA when the point of injection was the driven rod at point A. This highlights the significant effect of ionisation, since the driven rod at point A was 9 m long and was in 500 Ω m soil, whereas the rod at point B was 1,5 m long and was in 1000 Ω m soil. For current injection at point B, the remaining 28 kA splits between the two branches of the trench earth. However, a maximum current of 8 kA flows through the end of the trench (compared to 1,4 kA) into building A – this is because the earth electrode around building B is not as significant as that around building A.

5. CONSIDERATION OF A SUITABLE EXPERIMENTAL PROCEDURE

Although the model presented in this paper has been developed using a combination of elements for which the behaviour is well known [1,3], the behaviour of an earth electrode system as a whole

needs to be verified experimentally. Due to the difficulties involved in performing validity experiments, very few practical results are available for such a system. The following sources contain practical results [8,11,15–17], although much of this work was focused on the behaviour of transmission line earth electrodes. It is therefore proposed that a simplified full scale experiment be conducted to verify the model presented in this paper. Additionally the effect of wave-shape on ionisation in the soil and the modelling of surface flashover effects will be investigated.

The following general comments must be noted. Firstly, the impulse generator used must be capable of generating high currents of suitable wave-shape. However, due to ionisation effects, the impedance of the electrode under test will be not be as high as the low current value permitting the generation of higher currents. Secondly, the experimental soil conditions will be largely dependent on conditions near the impulse generator and the size of the experimental earth electrode system will also be limited by the space available near the generator. Therefore, it may be that only one driven rod with two trench earth connections can be tested. Finally, it is critical that all loop sizes must be minimised to avoid measurement errors due to coupling effects as well as to limit unnecessary loading of the generator by the increased inductance.

Table 2 : Impulse generator and experimental test parameters.

Parameter	Value
Number of stages	24
Per stage capacitance	0,75 μ F
Charging voltage	160 kV
Generator configuration	Series
Total circuit inductance	400 μ H
Load resistance	260–60 Ω

The low current resistance of a standard 1,5 m long, $\varnothing$ 12 mm rod electrode in 500 Ω m soil is 260 Ω –however even for currents as low as 5 kA this value will fall to as low as 60 Ω [1].

Assuming an impulse generator and test circuit with the characteristics summarised in *Table 2*, the predicted output current waveform can be calculated using state-space analysis. Two predicted current waveforms can be seen in *Figure 5* – the first is for a 260 Ω load and the second for a 60 Ω load. Note that the actual test waveform will be closer to the second due to ionisation effects although the two curves define the envelope for the current waveform. The significant negative overshoot will be ignored and the experimental results

only considered for the first 10 μs .

Note that a standard voltage impulse generator has been used in the above proposal where the capacitors are discharged in series. This is not an optimal configuration for current impulse generation. If possible from a practical point of view, it would be better to use a series-parallel combination of capacitors. Careful consideration is also required to arrange and verify a suitable measurement circuit configuration.

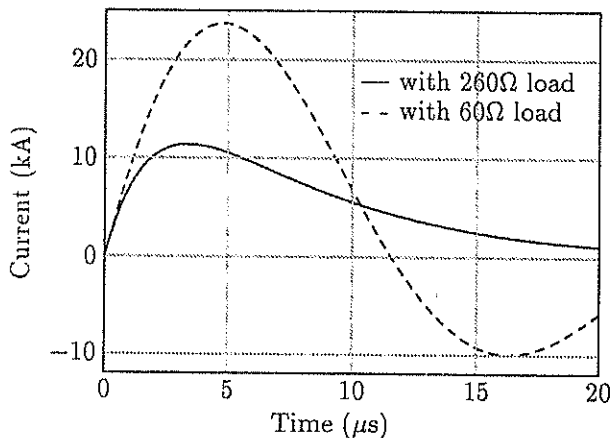


Figure 5 : Predicted experiment current waveforms

6. DISCUSSION

Using basic building blocks and modelling techniques, a suitable model for a site-wide earth electrode system has been developed and investigated with the ultimate intention of including the model in the analysis of larger systems. For example, the inclusion of other lightning protection system components and additional models such as the model for induced voltage on signal cables [2] would enable accurate quantification of surge protective device ratings for a site. It is important that the predictions of the complete earth electrode system model are tested experimentally, providing further insight into possible improvements of the model.

7. CONCLUSION

Breakdown within the soil surrounding an earth electrode results in significant additional lightning current dispersion. Using the model, it was shown that a relatively short driven rod subjected to a lightning transient current can achieve the performance of a substantially longer rod.

Mutual interaction between earth electrode system components should be included for elements reasonably close together (less than 10m).

Although the behaviour of the individual compo-

nents of the earth electrode system model is well understood, the need to verify the performance of the complete model using a full-scale experiment is identified.

8. ACKNOWLEDGEMENTS

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A NEW HIGH-POWER RESONANT INVERSION PULSER TOPOLOGY EMPLOYING REDUCED SEMICONDUCTOR RATINGS

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Abstract: High pulse energy- high pulse repetition rate pulsers are used, among others, to power Pulse-lasers and Pulsed Corona Reactors. These systems invariably use solid state switching on the low-voltage side and gas-filled thyatron switches on the high voltage side with pulse compression. New design trends favour switchmode power supplies for high voltage capacitor charging and replacement of the thyratrons with series semiconductor stacks. The author shows in this paper that resonant switching with semiconductor switches on the low voltage side is a viable alternative when voltage inversion is used on the low-voltage side and compression is carried out on both low- and high-voltage sides.

Keywords: Pulse Power, Resonant Energy Transfer, Pulse Power Supply, Electromagnetic Pulse Compressor.

INTRODUCTION

High energy, high pulse repetition rate pulse power systems, or *pulsers*, such as those used for pumping pulse lasers in isotope separation and for powering pulsed corona pollution control, require pulse voltage rise times below 200 ns, to be delivered to the load. In these applications, gas-filled thyratrons have held their ground for a long time. These switches are capable of switching peak currents in the kA regime and are capable of switching directly at several tens of kV. Unfortunately thyratrons have finite heater life times and these lifetimes are even more severely reduced by other factors when working at high pulse repetition rates and high peak currents. To overcome this difficulty in the more powerful systems, the trend in design is to replace thyratrons with semiconductor stacks on the high voltage side and to ease the switching load on these stacks by employing one or more stages of compression between the switch and the load.

An alternative design technique is to carry out the switching at low voltage, before transforming the pulses to the high voltage level demanded by the

load. This method has inherent difficulties. Firstly, the technique only becomes technically and economically viable when the pulse transformation after switching can be carried out in the region of 1 or 2 μ s. These short pulse widths are essential to limit the pulse transformer and pulse compressor magnetic volumes. In addition, shorter pulse widths of transfer inherently furnish less pulse time jitter, often a mandatory requirement with pulse laser pumping where outputs must be synchronised.

The topology described below has several integral features for overcoming the above difficulties by increasing the compression on the low-voltage side and employing voltage inversion there. Two versions of the topology are discussed in this paper. When the pulse output requires to be synchronised with others, and for high-frequency applications, the *Resonant Inversion Pulser* (RIP) is used and for high pulse power levels at relatively low rep-rates, with low semiconductor switching burdens, the *Super Inversion Pulser* (SIP) topology can be used. The above two topologies are discussed here and experimental results are shown for the first type, in which a prototype has been manufactured and tested.

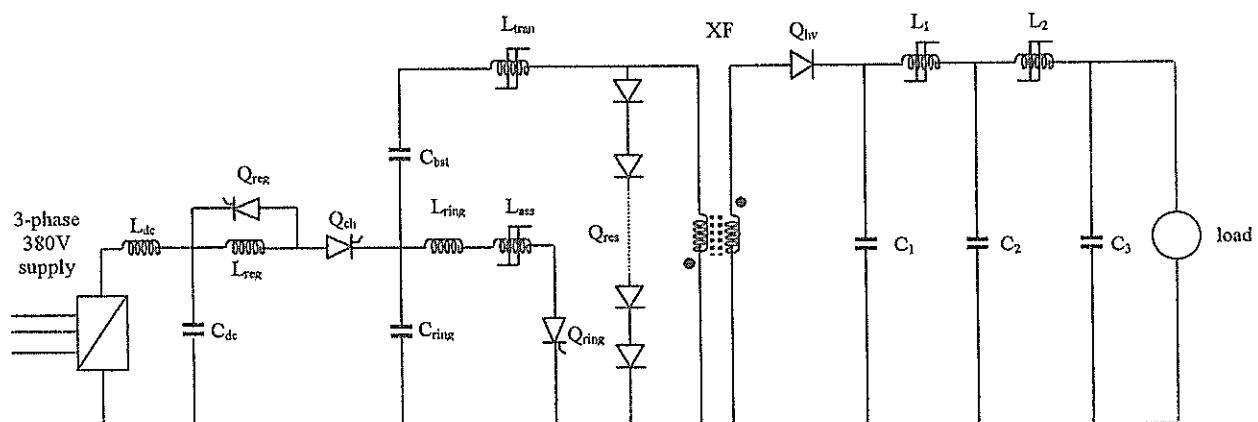


Figure 1 – Simplified power diagram for the Resonant Regulating Pulser topology

Unfortunately a full discussion of the mechanisms and techniques used in resonant energy transfer and solid state pulser operation lies beyond the scope of this paper and the reader must be referred to the literature^[1,2,3].

1. THE RESONANT INVERSION PULSER TOPOLOGY

A simplified diagram for the RIP power circuit is shown in Figure 1. The infinitely variable output DC-voltage from the Phase-controlled AC/DC converter is smoothed by L_{dc} and C_{dc} and is closed-loop controlled by the output pulse voltage requirement between 0 and 480 V. Initially the *primary transfer capacitor* combination C_{bst} and C_{ring} will be discharged. The *charging phase* is initiated when Q_{ch} turns on and C_{bst} and C_{ring} are resonantly charged from C_{dc} through L_{ch} to about 915 V each as shown in Figure 2.

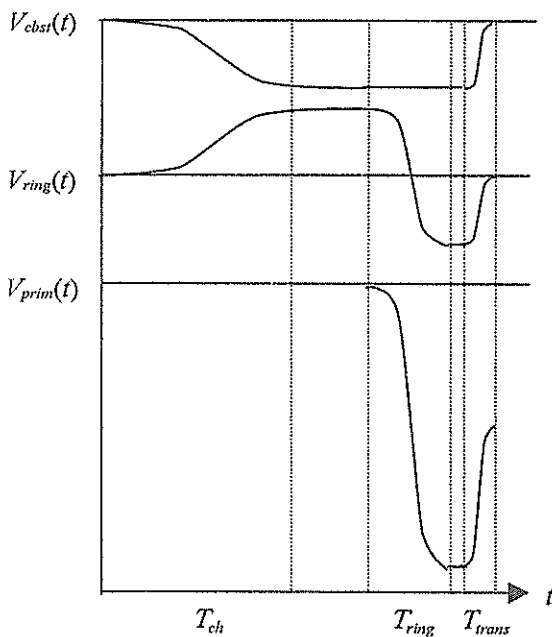


Figure 2 – Primary capacitor voltages

Capacitors C_{bst} and C_{ring} will respectively have co-sinusoidal charging voltage profiles $V_{cbst}(t)$ and $V_{ring}(t)$ as shown with pulse time T_{ch} in the order of 2.14 μ s. C_{ring} is charged directly and C_{bst} is charged through L_{tran} and reset-voltage clipper Q_{res} to permit a resetting voltage of about 7 V in the pulse transformer primary.

Pulse voltage de-queuing^[3] is optional, depending on the required output pulse voltage stability and time jitter specifications. De-queuing takes place by

triggering thyristor Q_{reg} when the voltage across C_{ring} crosses a pre-set threshold during the second half of the co-sinusoidal charging period. This immediately terminates the transfer of energy by force-commutating Q_{ch} and circulating the current through Q_{reg} and L_{reg} until Q_{ch} is triggered again. The de-queuing voltage waveform is not shown in Figure 2. If voltage de-queuing is not employed, Q_{ch} commutates naturally at the end of the charging pulse.

The *ringing phase* is initiated by triggering Q_{ring} , bringing about resonance in the circuit formed by C_{ring} , L_{ring} , L_{ass} and Q_{ring} , resulting in a co-sinusoidal voltage reversal in C_{ring} over the ringing pulse width T_{ring} of 7.73 μ s. The voltages across C_{ring} and C_{bst} add during charging and remains zero during charging but will erect a co-sinusoidal voltage profile with voltage $V_{prim}(t)$. This voltage will achieve a maximum value of 1712 V during ringing as shown in the bottom trace in Figure 2. The *magnetic assist inductor* L_{ass} boosts the di/dt capability of Q_{ring} , enabling this relatively small thyristor to switch a peak repetitive value of 1329 A at below 2 μ s at frequencies up to 1 kHz.

During the charging phase, the charging current through C_{bst} will fully reset the saturable magnetic core in L_{tran} . While voltage erection is taking place across C_{bst} - C_{ring} during ringing L_{tran} will initially be unsaturated and inhibit current flow. It is designed to saturate at the end of the ringing cycle, however, and the combined charge in C_{ring} and C_{bst} will be resonantly transferred through the pulse transformer XF and the high voltage diode D_{hv} to the first high voltage capacitor C_1 in the two-stage series pulse compressor chain. The *transfer pulse-width* T_{tran} is designed for 2.14 μ s. The delays shown between the three successive phases of *charging*, *ringing* and *transfer* ensure de-coupling between the circuits.

Table 1 – Comparison of alternative pulser designs

PULSER TYPE		R ² P ²	RIP	SIP
Output pulse-width to load	s	2.00E-07	2.00E-07	2.00E-07
Peak output pulse-current	A	952.28	952.32	952.32
DC Bus voltage	V	456.791	476.0243	505.1571
Primary pulse voltage	V	887.0777	913.2252	916.7149
Pulse-width to compressor	s	6.72E-06	2.15E-06	2.15E-06
Pulse-energy to compressor	J	2.74	2.50	2.50
Pulse-time of transfer	s	6.72E-06	2.15E-06	2.15E-06
Peak Primary Pulse Voltage	V	887.0777	1712.47	1719.013
Peak Primary Pulse Current	A	1647.029	2278.866	2539.745
Switched-current pulse-width	s	6.72E-06	7.73E-06	3.61E-05

Peak switched Current	A	1647.029	1329.87	322.4961
Effective switched Current	A	95.40679	82.65866	43.33871
Average switched Current	A	7.037759	6.542851	7.415716
Rate of rise switched Current	A/ us	820.5673	505.6018	27.17354

A two-stage series pulse compressor is employed as shown in Figure 1^[4] to compress the half-sinusoidal current pulses, delivered by the power supply, from 2.15 μ s down to the required 200ns required by the load.

The merits of the RIP topology only begin to emerge when it is compared with that of the Resonant Regulating Pulse Power (R^2P^2) topology that is used to this day^[4]. Table 1 lists the performance of the RIP topology against the R^2P^2 topology for a 2 J, 33 kV and 1 kHz specification.

The RIP boasts two inherent steps of voltage magnification that enables higher primary pulse voltages to be attained without additional transformation of the supply source voltage. The rectified DC voltage is virtually doubled from 476 to 913 V during charging and then again to 1712 V by voltage inversion. In addition, the RIP permits this high primary voltage to be used, requiring the main semiconductor switch to be rated only for 913 V.

The pulse-width of transfer in the RIP topology is seen to be 2.15 μ s, compared to 6.72 μ s in the older

design. This shorter pulse width brings about a commensurate saving in pulse transformer and pulse compressor magnetic volume and pulse time jitter is also proportionally reduced. The peak repetitive current is also seen to be down from 1647 A to 1329 A, in spite of the shorter pulse width.

The enlargement of inductors on the LV side and the distribution of compression between HV and LV require careful consideration and design. It was shown by the author that compression at LV is more economical than at HV because of a higher utilisation^[1] of the saturable cores. In addition, less expensive magnetic core materials can be used at LV than at HV.

2. THE SUPER INVERSION PULSER TOPOLOGY

The SIP Topology shown in Figure 3 requires an even lower semiconductor current rating than the Inversion Pulser Topology depicted in Figure 2. In this circuit, a saturable inductor L_{ring} replaces the ringing thyristor and the charging thyristor Q_{ch} now performs the active switching. It is necessary to actively reset the ringing inductor and this is accomplished through a current-source and reset winding (a constant voltage source with series inductor can be used in practice).

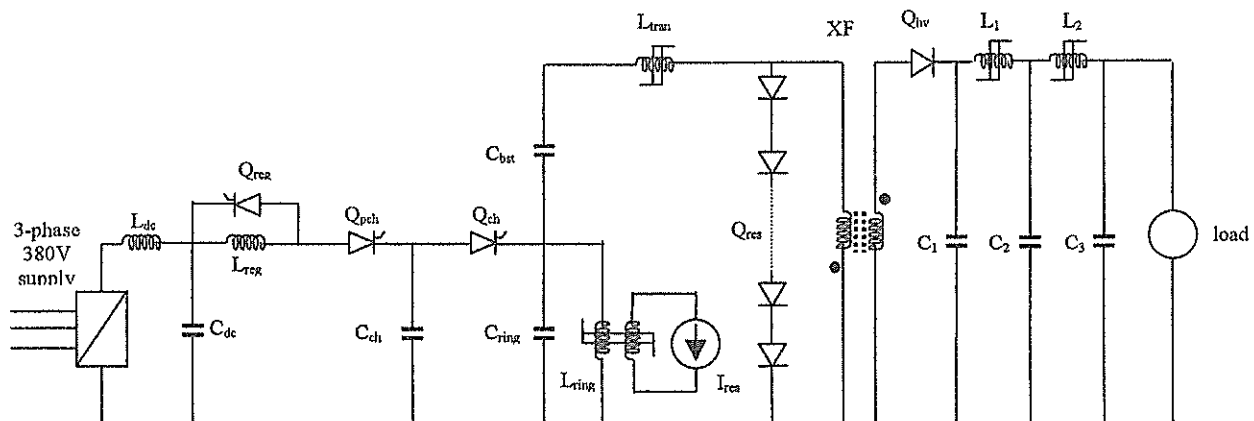


Figure 3 – The Super Inversion Pulser Topology

Operation is as follows: The pre-charge capacitor C_{pch} is resonantly charged, with- or without de-queuing as before from C_{dc} by triggering Q_{pch} . As soon as transfer is complete in the pre-charging loop, Q_{ch} is triggered and transfer of energy then takes place resonantly from C_{ch} to C_{ring} and C_{bst} as before. The co-sinusoidal rise of voltage across C_{ring} must be completed before inductor L_{ring} saturates and resonantly reverses the polarity across C_{ring} . L_{ring} is

brought out of saturation by the current through it reversing, as in the case of the thyristor and “commutates” with similar results. Voltage erection takes place over C_{ring} - C_{bst} as before. It must be noted here that where ringing could be initiated by actively triggering Q_{ring} in the RIP, ringing is now initiated as a function of the volt-seconds area under the charging voltage profile across C_{ring} . Because the hold-off of L_{ring} is, among others, a function of its core volume,

the charging pulse time in the preceding loop will be limited through that from an economic point of view.

The pulse-width of charging can be adjusted to be many orders larger than that of ringing but must still be accommodated between successive pulses. This can become a problem at high pulse rates because C_{pch} must be fully charged before the charging cycle can commence. One way of overcoming this difficulty is to install several front-ends and several capacitors C_{pch} and to multiplex charging from these different capacitors C_{pch} while the others are in the process of charging.

The performance of a 2 J SIP, employing an identical compressor to that used in the RIP unit, is depicted in column 5 of Table 1. Note that the peak current through Q_{ch} (the main semiconductor switch) is now only 322 A compared to that of 1329 A in Q_{ring} (the main semiconductor switch) in the case of the RIP unit. This reduction is achieved by virtue of the large pulse width difference of 7.7 μ s to 36 μ s.

3. CONSTRUCTION

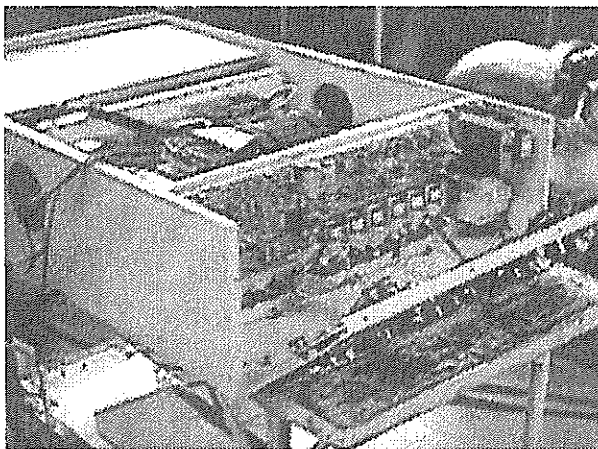


Figure 4 – Internal arrangement of Resonant Inversion Pulser

A prototype of the Resonant Inversion Pulser was constructed and tested at low frequencies without spending any time in tuning or optimising it at this stage. Although this pulser is equipped with the facility for active de-queuing regulation, the relevant circuits have not been commissioned yet at the time of writing. The prototype pulser is shown opened in Figure 4 and with its cover installed in Figure 5.

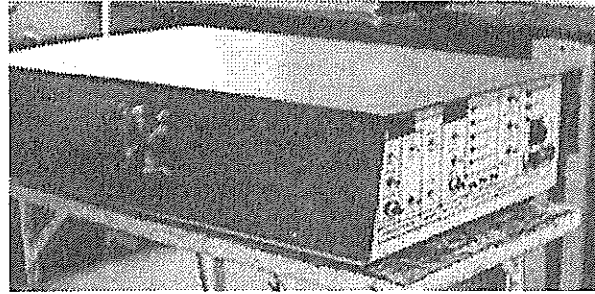


Figure 5 – Assembled Resonant Inversion Pulser

The pulser is sheet-steel housed and has three basic compartments respectively containing the control section, the low voltage section and the oil-filled high voltage section. As shown in Figure 4, the front control panel is hinged to furnish access to the control section.

4. EXPERIMENTAL EVALUATION

The pulser was connected to a CO₂ Transversely Excited Atmospheric Laser and was also tested whilst driving a Pulsed Corona Reactor. The waveforms shown below relate to the latter for an operating frequency of 50 Hz.

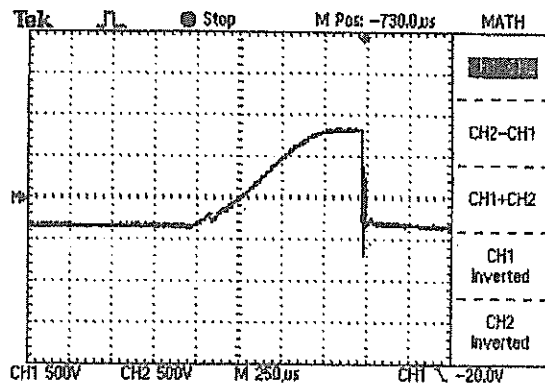


Figure 6 – Expanded waveform of charging and ringing voltage across capacitor C_{ring}

The oscillogram in Figure 6 shows the capacitor C_{ring} voltage collectively over the charging, ringing and transfer periods. Charging is seen to take place co-sinusoidally in about 800 μ s from an initial voltage value of about -350 V to about 800 V. Then there is a delay of about 150 μ s before ringing is initiated.

The ringing and transfer voltage profiles across C_{ring} is expanded in the oscillogram in Figure 7.

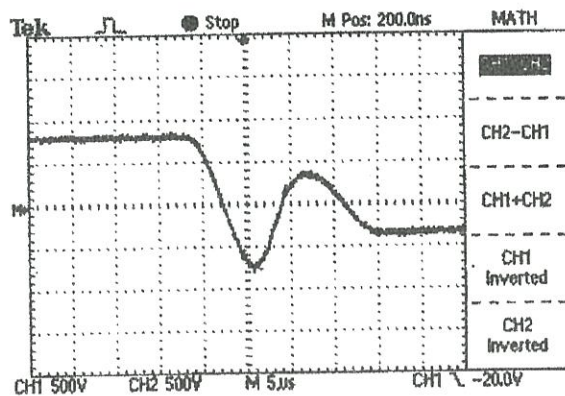


Figure 7 - C_{ring} voltage waveform during Ringing and transfer

In this oscilloscope, the voltage across C_{ring} is seen to reverse from the charged value of 800 V to a negative value also close to -800 V in roughly 5.5 μ s and conforms to the basic design criteria. Transfer then takes place in roughly 4.5 μ s, but instead of reducing the C_{ring} voltage to 0 V, it over-swings and then oscillates back to the initially observed -350 V level. Although the negative bias across Q_{ring} assists in commutation and in preventing accidental triggering, it has other unwanted implications and must be brought down to a more desirable level of about 50 V.

The cause of the above behaviour is primarily a too-large leakage reactance in the pulse transformer and as a result of that a too-small holdoff in the transfer inductor L_{tran} . Redesigning the pulse transformer and replacing the ferrite core with one of the newer high-flux excursion cores now available on the market will rectify the problem.

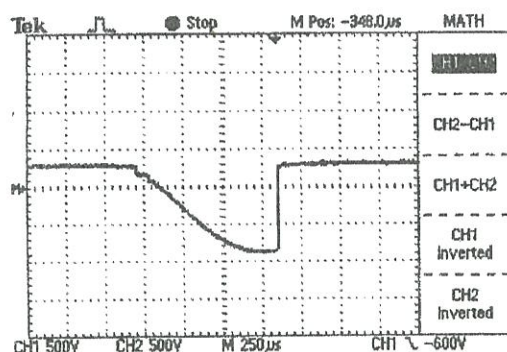


Figure 8 - Capacitor C_{bst} voltage

The expanded C_{bst} voltage is shown in Figure 8 in which it begins at +350V and then falls to zero again. This voltage is the voltage across C_{ring} during this time. The voltage across C_{ring} is zero

between pulses. The peak magnitude of the C_{bst} voltage is seen to attain the level of 800 V, which is also that of C_{ring} . The output voltage waveform from the pulse transformer is shown in the oscilloscope in Figure 9.

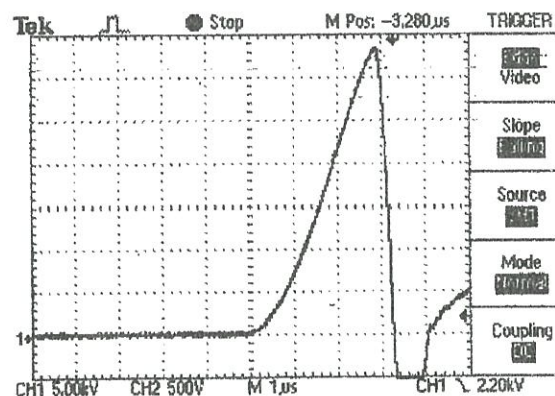


Figure 9 - Pulse Transformer output voltage

The voltage waveform is seen to rise co-sinusoidally in 2.8 μ s from 0 to 38 kV but it is terminated at the upper side, before transfer is complete, by the premature saturation of Inductor L_1 . The designed transfer time is for 2.147 J and it is clear that the pulse width of the transfer pulse is too large. Note how this waveform now falls from 38 kV and over-swings, indicating resonant transfer of energy from C_1 to C_2 . The observed pulse width of transfer indicated by this declining part of the C_1 voltage, corresponds with the designed pulse-width of transfer of 500 ns in the second stage of the compressor. The over-swing is partially the result of the premature saturation of L_1 and partly due to reflections from the plasma reactor load that has not been well matched yet when the shown waveforms were recorded.

The plasma reactor circuit superimposes the output pulses of the compressor on a DC-supply. The necessary circuitry to implement this has not been completed yet and the matching between it and the compressor output is poor without it and produces unwanted oscillations so that that part of the results will not be presented in this paper.

5. CONCLUSIONS

Although there is still a discrepancy between the designed and achieved transfer pulse width due to a miscalculation of the pulse transformer leakage inductance, the pulser is performing in accordance with its design. Because of the limited scope of this paper, it is not possible to discuss the intricacies of the design but suffice it to remark that the ferrite core material used in the construction of the RIP is very

unsuitable for that application. This material imposes a severe restriction on the compression that can be achieved because of the limited flux-excursion that ferrites can furnish and it will have to be replaced with one of the newer crystalline or amorphous magnetic materials. Further work also has to be carried out to commission the de-queuing regulation and to complete the load DC-pulse superposition circuit.

A complete design has been carried out for the Super Inversion Pulser, but construction of this pulser is now dependent upon additional funding.

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DEVELOPMENT OF A 45 KVA UNIVERSAL SEMICONDUCTOR ELECTRIFICATION DEVICE

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Abstract: The application of constant voltage source converters in cost-effective electrification configurations is discussed in this paper. The main purposes of this technology are as follows: to dynamically regulate voltage in a network and to convert power between multiphase and multi-frequency systems. This paper addresses possible application opportunities and shows simulation and experimental results.

1. Introduction

The development of this device started with the idea of applying constant voltage source (VS) converters for cost-effective electrification of rural communities, i.e. the USE (Universal Semiconductor Electrification) concept [10]. The USE device has the inherent capability of compensating for large voltage drops, voltage variations and poor power quality. In remote electrification schemes these attributes can also be used to compensate for a sometimes widely varying supply voltage and can interconnect multiphase, multi-frequency and distributed generation systems. Modularity in the system makes it easy to adapt to a specific need. The functions of the USE device can broadly be classified as voltage regulation, power conversion and power conditioning.

To illustrate these functions a hypothetical single-wire-earth-return (SWER) electrification project is discussed. A line feeds a rural community 30 km away from the nearest MV power line as shown in Fig. 1. The total community power consumption is approximately 150 kW and 10 kVAR (lagging). The load consists of one commercial farm (D) and three residential areas (A-C). Each residential area consists of 6 houses with maximum demands of 5 kVA, i.e. a maximum single-phase demand of 30 kVA per area. The farm demands a single-phase supply of 5 kVA and a maximum three-phase supply of 40 kVA. The project philosophy was to electrify the community with minimum capital investment.

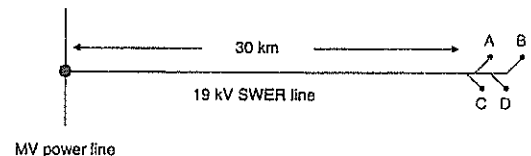


Fig. 1: The application of the USE device on SWER lines

The capital cost of a SWER line is approximately 35% less than the conventional three-phase option [5]. Due to the low demand from this community and uncertainty in future growth, it is decided to reduce the cost further by using steel wire (stay wire). The cost of steel wire is approximately R 120 per km. Bantam conductor costs R 690 per km. The span length of steel wire is 2 times that of Bantam. These considerations cause an added cost reduction of approximately 30%. Thus the line cost can be reduced to approximately 35% of the conventional cost, i.e. the line cost is reduced from R 435 000 (22 kV Bantam, 3-phase) to R 152 000 (19 kV steel, SWER). An estimated amount of R 283 000 is saved. However, poor line quality (electrical characteristics) cause poor voltage regulation and poor power quality. A three-phase supply is also not available to the customer. These issues are addressed by using 4, 45 kVA USE devices on the LV side of the SWER transformers. The cost of these devices is estimated at R 180 000. The cost saving is estimated at R 103 000.

Voltage Regulation

Conventionally, voltages on distribution lines are boosted by incorporating auto-transformers [6] and tap-changers at the MV levels. Fig. 2 shows the equivalent circuit diagram of such a device.

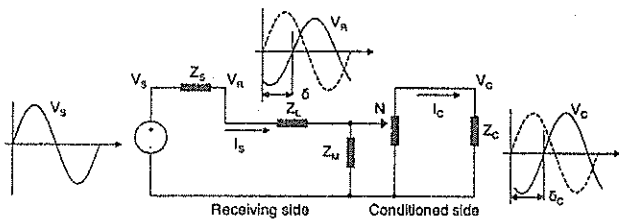


Fig. 2: Equivalent circuit diagram of an auto-transformer voltage regulator

The leakage impedance Z_L adds to the line impedance and can thus cause an additional voltage drop that has to be compensated for. A mechanically controlled tap changing mechanism changes the turn ratio N . Thus, regulation increases the phase shift (δ_c) in the compensated voltage.

The USE device can perform voltage regulation on the LV side of the distribution system. It controls the voltage in real time and frequent changes in load or system voltage does not influence the conditioned voltage V_c . The equivalent circuit of the device is shown in Fig. 3. The receiving side is equivalent to a constant current source and the conditioned side is equivalent to a constant voltage source (VS). The two sides consist of VS converters connected in a back-to-back configuration.

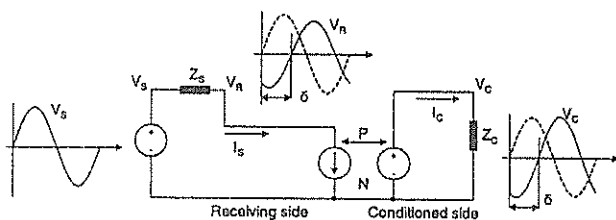


Fig. 3: Single line diagram illustrating voltage regulation with the USE converter

It is important to realize that for both voltage regulation devices the receiving end current must increase to boost the voltage so as to sustain the same power demand (P) on the conditioned side. An auto-transformer also has to transfer the reactive power required by the load, but the USE converter isolates the load's reactive power demand from the receiving side. The LV line current required for regulating the voltage to 240 V on a 19 kV SWER system can be calculated as follows.

$$V_R = V_S - I_S Z_S$$

$$(\times I_S) \Rightarrow V_R I_S = P = V_S I_S - I_S^2 Z_S$$

$$\Rightarrow I_S = \frac{V_S - \sqrt{V_S^2 - 4 Z_S P}}{2 Z_S}$$

The line's capacitance to ground is neglected and only series inductance or resistance is considered as part of the line impedance. (In practice, however, the effect of capacitance might not always be negligible.) Equation (2.1) is plotted for the major SWER line types in Fig. 4. This figure also shows lines representing the thermal current limit of the lines transformed to the LV side.

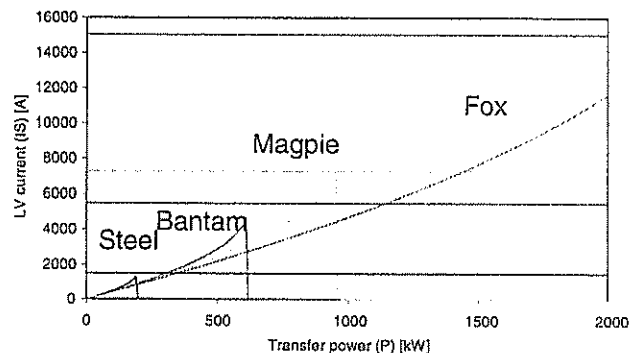


Fig. 4: Input current versus power transfer of the USE converter

The increased line current, drawn to maintain the load power, causes an increased voltage drop across the line. Due to this the conditioned voltage can only be regulated from a receiving end voltage of 50% the nominal voltage and higher. From Fig. 4, the maximum permissible power demand is obtained as approximately 200 kW for this 30 km SWER line. The line is used just below its thermal rating. In Fig. 5 the normalized voltage drop is plotted against power transfer. These graphs can be used to obtain the additional voltage drop when the load increases.

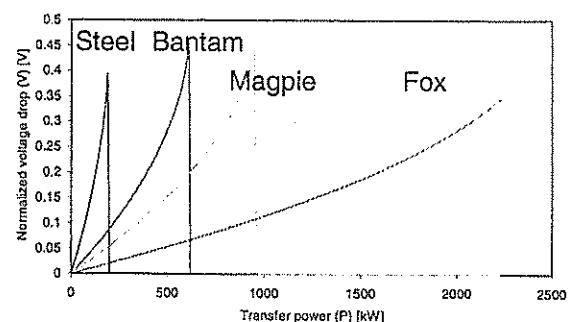


Fig. 5: Normalized voltage drop versus transfer power

If the maximum load demand should increase, an additional strand could be added to the existing conductor to strengthen the line to the desired power level. Three 45 kVA single-phase USE devices are used for supplying the three residential areas with regulated power. The farm, however, needs 40 kVA three-phase power.

Power Conversion

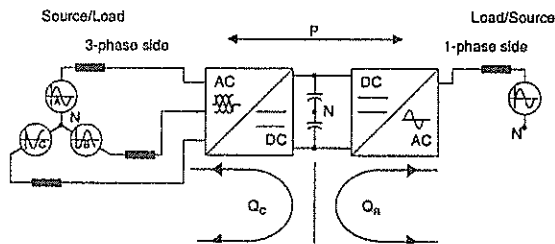


Fig. 6: Three-phase-single-phase USE device

The USE device (three-phase-single-phase version in Fig. 6) operates by converting power at the receiving end (the right side) to dc. The dc voltage can again be converted into single-phase or three-phase ac. The dc voltage is boosted to levels up to 800 V, well above the value required to regulate 400 V line-to-line voltages in a multi-phase system. Three-phase power is normally used to drive induction machines. Three to four times nominal current is needed during start-up for a limited amount of cycles. A comparison can be made between the USE converter and the rotary converter currently on the market for generating three-phase. Not much literature could be found on the performance of this device, but it seems that the high start-up current of the motor influences its rating. The three-phase voltage waveform phase shifts are a function of current and thus vary with load [12]. The USE converter regulates three-phase voltages independent of the current drawn by the load. It will be shown in the next paragraph that the peak rating of power electronics is well above the average rating that makes it ideal for applications requiring occasional peaks.

Power Conditioning

Due to the intermediate power transfer link, the USE device can temporarily store energy to supply it again in the next half cycle. In this way it can compensate for both leading and lagging reactive power. Thereby the two sides of the USE device

are isolated in terms of reactive power as illustrated in Fig. 6. Poor power factor, e.g. during motor start-up, thus has no influence on the line. However it places additional constraints on the power rating of the device as discussed in the following paragraph.

2. The Device Configuration

The core building blocks of any power electronic converter are the switching or modulating bridges between ac and dc systems. These bridges are commonly known as phase-arms. A single-phase-arm is shown in Fig. 7. For each additional phase required, on the source side or load side of the device, an extra phase-arm is added. Filters on the output of phase-arms isolate high frequency components, i.e. multiples of the pulse-width modulating (PWM) switching frequency.

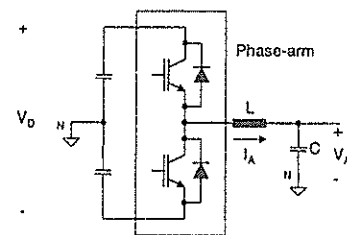


Fig. 7: Single phase-arm with filter

Apparent power (S) flow between the ac side and dc side of phase-arms depends on the phase relationship of the fundamental component of the synthesized current and voltage I_A and V_A . When the net flow of active power (P) is from the ac side to the dc side the phase-arm acts as a rectifier and can boost the dc voltage to a required value. When the net flow of P is from the dc side to the ac side the phase-arm acts as an inverter regulating the ac voltage to a required value. Reactive power (Q) can be superimposed on the active power component. When the ac current leads the ac voltage by between 0° and 90° Q is leading and when the ac current lags the ac voltage by between 0° and 90° Q is lagging. A single-phase source connected between two passive phase-arms (full-bridge configuration) will provide a maximum dc link voltage of

$$V_{d\max} = \sqrt{2}V_R$$

V_R is the line voltage. A two- or three-phase source connected between phase-arms will provide a maximum dc link voltage of

$$V_{d\max} = \sqrt{2}V_{RLL}$$

V_{RLL} is the line-to-line voltage. By switching the phase-arms, V_d can be boosted to provide a high enough voltage to be used by the inverting stage for synthesizing line voltages or line-to-line voltages to values determined by the current ratings of the lines, transformers and phase-arms. For a single-phase to single-phase USE converter drawing only active power, the rectifier rating depends on the receiving voltage and load power as follows

$$\begin{aligned} V_R I_R &= P \\ \Rightarrow I_R &= \frac{P}{V_R} \end{aligned}$$

If the device is required solely to compensate for reactive power on the rectifier side, the current rating of the rectifier stage would be increased

$$I_R = \frac{Q_R}{V_R}$$

This does not influence the inverter stage rating of the converter, but places an additional constraint on the size of the capacitors of the intermediate power transfer link. Thus the current rating of the rectifier side is

$$I_R = \sqrt{\left(\frac{P}{V_R}\right)^2 + \left(\frac{Q_R}{V_R}\right)^2}$$

Similarly the rating of the inverter stage can be calculated as

$$I_C = \sqrt{\left(\frac{P}{V_C}\right)^2 + \left(\frac{Q_C}{V_C}\right)^2}$$

The maximum rms power rating of the device is limited by the thermal rating of the power components. Superimposed on the load current is a high frequency current that is filtered by the low-pass filter. This current has a maximum value of 20% of the load current at full load and maximum dc link voltage. The maximum 3-phase apparent power rating at full voltage is

$$S_{3P} = \sqrt{3}I_{3P\max}V_{3P\max} = 110\text{kVA}$$

Similarly the maximum 2-phase apparent power rating at full voltage is:

$$S_{2P} = I_{2P\max}V_{2P\max} = 240 \times 400 = 96\text{kVA}$$

The maximum 1-phase apparent power rating:

$$S_{1P} = I_{2P\max}V_{2P\max} = 240 \times 230 = 55.2\text{kVA}$$

The component ratings further show that the silicon can withstand occasional current surges to approximately 5 times the peak current ratings. The repetitive peak current rating makes provision for current overshoot due to diode reverse recovery and transients during switching.

3. Cost of the USE Device

The cost of the USE device is approximately R 1000 per kVA. This is approximately 10 times the cost of auto-transformer voltage regulators. The main reason is that current IGBT technology can withstand voltages of up to 3.3 kV and need to be used on the LV side. The high LV side current rating makes these devices difficult to compete with regulators on the MV side. However, one has to weigh the benefit of real time control against the increased cost. The USE device will automatically compensate for power quality problems. For outages, however, some kind of energy storage equipment should be connected to the dc link, e.g. batteries or flywheels.

The USE converter can be a major competitor for the rotary converters. The main reason is that the device is ideal for supplying high peak currents during motor start-up. When comparing the cost per kW load, the costs of these two devices are approximately the same. The benefits of voltage regulation and power conditioning with the USE converter are then added as a bonus.

4. Simulated and Experimental Results

Fig. 8 shows the output voltage waveforms and the output current, of a phase, obtained with the three-phase USE inverter. This result was obtained through the space-vector control method.

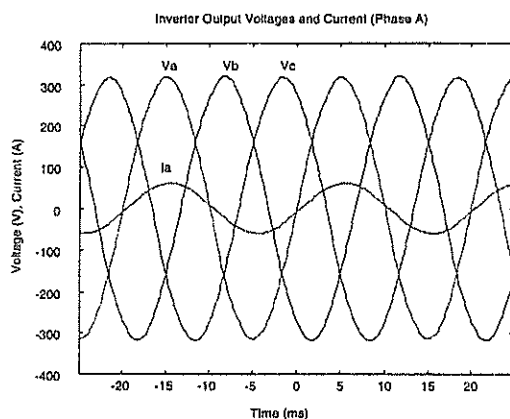


Fig. 8: USE inverter output voltages and current

Fig. 9 shows changes in current drawn by the rectifier side of a USE device. It can easily change from lagging to leading.

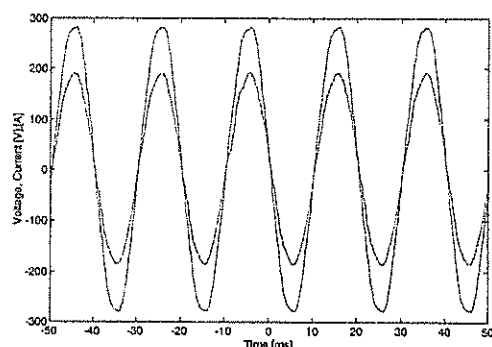


Fig. 9: Single-phase rectifier waveforms

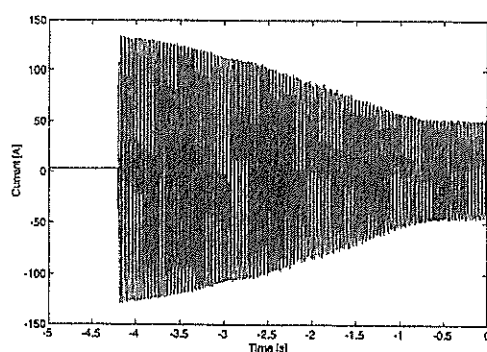


Fig. 10: Induction motor start-up

5. Conclusions

To conclude, typical applications for the USE device are listed below:

- Rural electrification using long radial lines

- SWER lines
- Sensitive loads
- Low power-quality lines
- Integration of distributed resources
- Uninterruptable power supplies
- Dual-phase distribution
- DC distribution

Various options exist for power conversion using the USE device, i.e. 2-phase to 3-phase, 1-phase to 3-phase, 3-phase to 1-phase, 1-phase to 1-phase, dc to 1-phase, dc to 3-phase, 1-phase to dc, 3-phase to dc. Bi-directional power is possible due to static switching of both sides. The percentage voltage regulation allowed at the receiving side, while still maintaining correct customer voltage levels, depends on system parameters, but can be up to 49%. By controlling the current-voltage phase relationship, compensation for load impedance (power-factor control) and line impedance through the USE device is possible.

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Boost Rectifier with a Neural Network Current Controller

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Abstract - This paper investigates the use of a continuously on-line trained Artificial Neural Network Current Controller for a boost rectifier, to determine the viability for its use in Variable Speed Drives, to overcome voltage dips on the mains supply. The paper analyzes simulation results of a Continuous on-line Trained Artificial Neural Network controller for a Variable Speed drive System, and shows that this current controller is dependent on the magnitude and frequency of the supply or load currents, and the Artificial Neural Network's internal variables. The practical results will show that an Artificial Neural Network current controller can be used with a boost rectifier, to overcome mains voltage dips. Finally the paper will indicate the advantages and disadvantages of a Neural Network current controller as opposed to conventional controllers such as PID or robust controllers.

Keywords - Neural Networks, Variable Speed drives, Current controllers, Boost rectifiers

1. INTRODUCTION

Most AC variable speed drives employ the configuration, shown in Fig. 1.

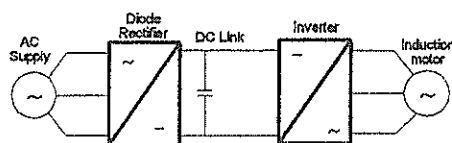


Fig. 1. Voltage-Sourced Inverter

AC Variable Speed Drives (VSD) are susceptible to voltage dips present in the mains supply, caused by a change of input impedance due to either a mains fault or mains disturbance, and results in a reduction in the control capabilities of the variable speed drive (VSD).

It is proposed to use a Boost rectifier as the front end rectifier for the VSD, to try and maintain a constant DC link voltage under limited fault conditions. Under fault conditions the supply impedance changes, therefore the current controller used in the Boost rectifier must be able to account for such changes in the supply impedance to ensure stable operation. Conventional current controllers suffer from the problem, in that they are tuned by pole placement, thus if any of the system parameters change, the controller becomes de-tuned, and experiences a loss in performance.

The Boost rectifiers input impedance is made up of two components (shown in Fig. 5), namely:

- 1) Line impedance, L - fixed impedance
- 2) Supply impedance, L_s - variable impedance

It is the varying supply impedance, L_s , which causes mains voltage dips to occur, thus the controller must be able to identify the changing impedance in order to reduce the effects of voltage dips. An investigation to determine the feasibility of using a ANN current controller for a Boost Rectifier and exploiting the self learning properties of the neural network to compensate for changes in the line impedance, and allow the controller to continuously adapt to changes in the supply impedance, thereby allowing for better stability, and for a more reliable system under fault conditions.

Section 2 provides a brief outline of the Boost Rectifier

configuration to be implemented with the ANN current controller.

Section 3 presents the findings of simulations performed for an ANN current controller for a Voltage Source Inverter (VSI) fed Squirrel Cage Induction Motor (SCIM) drive. These initial simulations will provide familiarity with ANN controllers, and an understanding of their stable areas of operation.

Section 4 and 5 will provide an understanding of the previously discussed ANN's conventions, and method that is required to obtain unity power factor operation of an ANN current controller.

Section 6 will analyze simulation results obtained for a ANN current controlled boost rectifier, and theoretical confirmation is obtained for the viability of utilizing an ANN current controlled boost rectifier, to overcome voltage dips.

Section 7 and 8 will describe the practical realisation of the COT ANN current controller, and present results to verify the viability of using a COT ANN current controlled Boost rectifier to achieve a regulated output voltage under fault conditions.

The results presented in this paper are unique, as other similar systems, [1], [2], have never highlighted these problems, whether off-line or on-line trained.

2. SYSTEM OUTLINE

To compensate for voltage dips, it is possible to use a boost rectifier in place of the diode rectifier, to achieve a constant DC link voltage. The COT ANN is used solely for current control, and a separate proportional or proportional/Integral controller is used to control the DC link voltage as shown in Fig. 2.

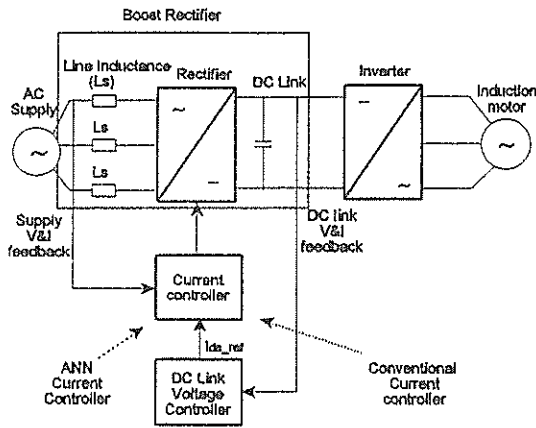


Fig. 2. Classified control loops

Before the ANN current controller for a Boost rectifier was investigated, a comprehensive study on an existing system [2] was first conducted, to gain an idea of the nature and form of the COT ANN as well as its operation under the non-linear effects of a VSI fed SCIM drive.

3. INVESTIGATION OF THE PERFORMANCE OF A COT ANN APPLIED TO A VSI FED SCIM

The first objective was to obtain familiarity with COT ANN current controllers and to evaluate under what conditions the ANN could operate and be stable, as previous work [2] neglected the non-linear effects of the inverter, which was found to cause the system to become unstable at high motor frequencies. Fig. 3 shows a block diagram of the simulated system.

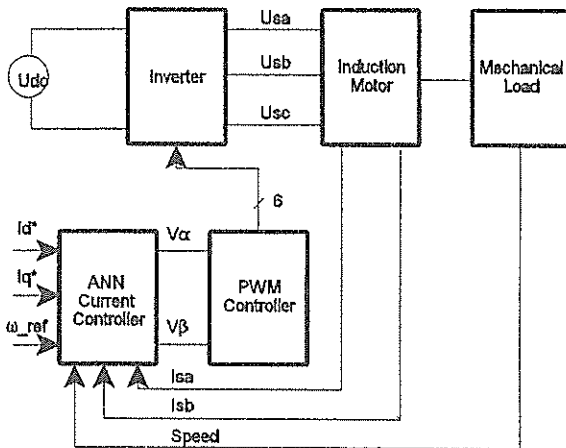


Fig. 3. CASED simulation Block

When the PWM controller and inverter blocks were added, the ANN employed in [2] became unstable for frequencies greater than 2 Hz.

Successive simulations results revealed that the transient response and tracking capability of the ANN is dependent on a number of factors [3], namely:

- 1) Motor Current magnitude,
- 2) Motor Frequency,
- 3) ANN learning rate (B),
- 4) Voltage constant (Cv),
- 5) ANN sampling frequency (fs),

4 ANN CONTROL LOOP DERIVATION

The ANN in section 3 was designed as a current controller for a VSI fed SCIM, but the same configuration can be used for the Boost rectifier, by making some minor changes in sign conventions. When controlling a VSI fed SCIM, the sign conventions conform to those in the "Motoring" block in Fig. 11, but when controlling a Boost rectifier, the conventions conform to those in the "Regeneration" block.

Thus to adapt the existing ANN current controller to the Boost rectifier, the internal feedback loops need to be altered to this new convention as if controlling a regenerative SCIM.

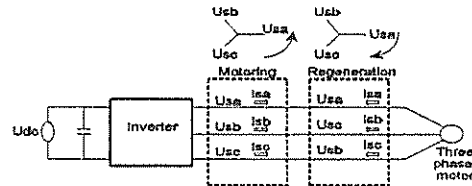


Fig. 4. Phase conventions for a three phase inverter

5. MAINS SYNCHRONIZATION

The ANN current controller must be synchronized to the mains supply, to achieve unity power factor operation, this is performed by inputting the phase to neutral voltages to the ANN and using them to synchronize to the mains. Fig. 5 shows a Conventional controller block diagram [6], and indicates the alterations that are required, so that the ANN may operate as the primary controller.

As the ANN operates in the Alpha/Beta reference frame [4], so all the inputs must be transformed utilizing the Power-Invariant Park's transform, where θ is set to zero, as follows:

Voltage equations:

$$E_{\alpha mains} = (U_{sa} - \frac{1}{2} * U_{sb} - \frac{1}{2} * U_{sc}) / \sqrt{3} \quad (1)$$

$$E_{\beta mains} = (U_{sb} - U_{sc}) / \sqrt{2} \quad (2)$$

The alpha/beta voltages are inputted to the ANN, and used to calculate the per unit mains Sine and Cosine components as shown in equation (3) and (4).

$$\cos = E_{\alpha mains} / M_{ag} U_s \quad (3)$$

$$\sin = E_{\beta mains} / M_{ag} U_s \quad (4)$$

Line current references are created which are in phase with the phase to neutral voltages of the supply, by using the sine and cosine components as follows:

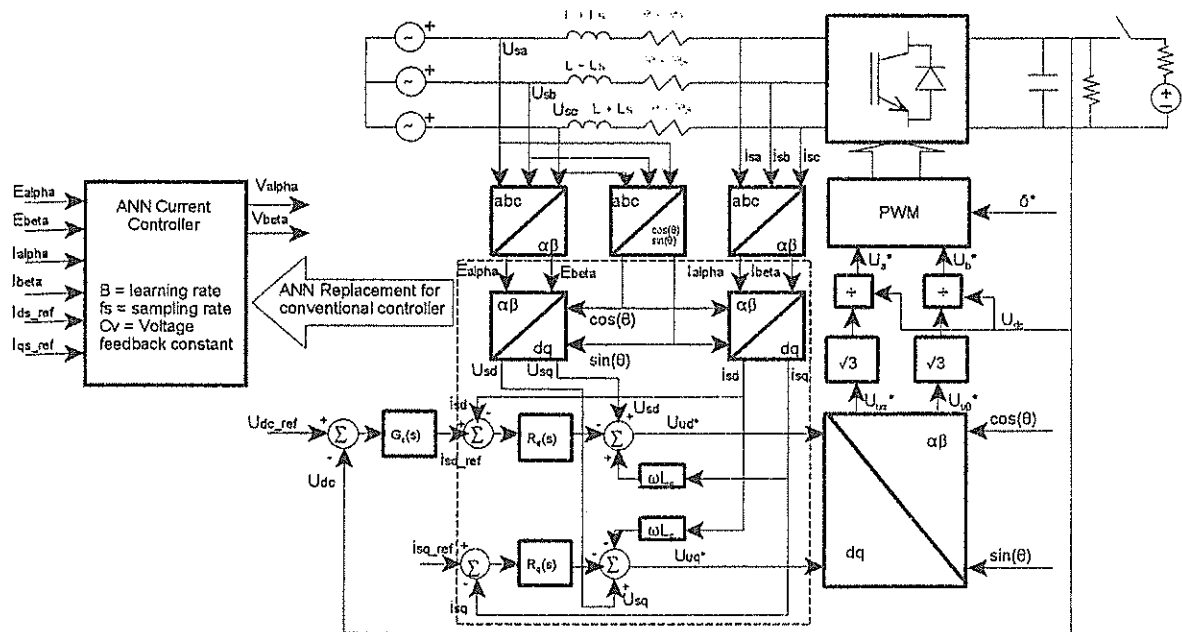


Fig. 5. Converting a conventional current controller to a COT ANN current controller

$$I_{a\_ref} = I_{ds\_ref} * \cos(\theta) - I_{qs\_ref} * \sin(\theta) \quad (5)$$

$$I_{b\_ref} = I_{ds\_ref} * \sin(\theta) + I_{qs\_ref} * \cos(\theta) \quad (6)$$

The ANN utilizes these reference currents, to obtain unity power factor operation (i.e. synchronous operation).

6. INVESTIGATION OF A COT ANN CURRENT CONTROLLER FOR A BOOST RECTIFIER SYSTEM

CASD user models were developed for system as shown in Fig. 6. The COT ANN's structure from the previous section had to be changed (more inputs were used, and the feedback sign conventions were altered as discussed above), so that it could control a Boost Rectifier, thus the ANN in this section is fundamentally different to the ANN used in

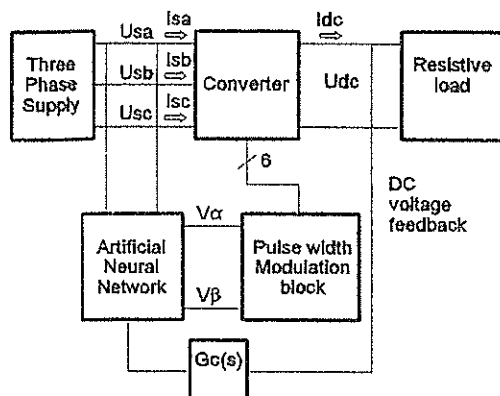


Fig. 6. System Block Diagram

the motor controller.

Simulations were performed to determine whether an ANN current controller could be used to control a Boost Rectifier, to minimize the effects of voltage dips.

In the system, the DC link Capacitor was pre-charged to its nominal voltage, allowing the controller to take control of the system faster, and to prevent the problem of inrush current to the Rectifier, Fig. 7 is an example of unstable operation of the ANN current controller, here the ANN demands very large currents which the Boost Rectifier can't source, hence the DC link voltage drops to zero, and the ANN loses control.

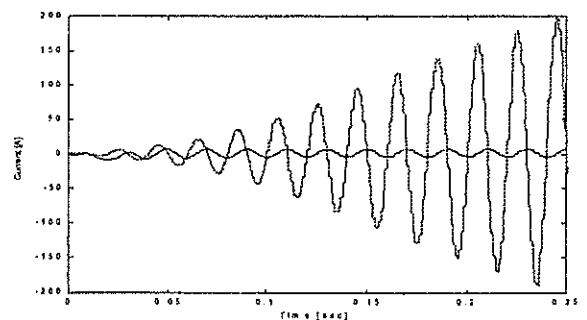


Fig. 7. Phase A supply current and ANN alpha current

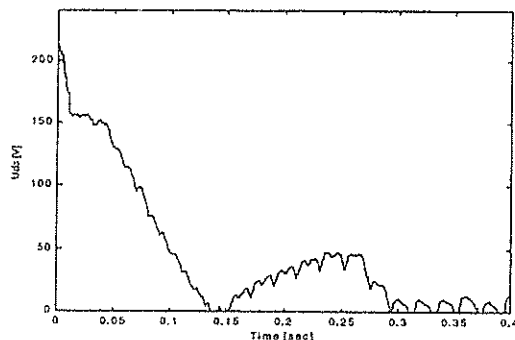


Fig. 8. DC Link voltage

The simulation results shown above illustrate instability of the ANN current controller when the incoming line impedance is too large. This represents the condition when there is a fault, and the Boost Rectifier is unable to source sufficient current to boost the DC link voltage.

The line inductance was adjusted (set to 10 mH), until the ANN current controller operated successfully, and then a number of successive tests were performed, to determine desirable operating point for the ANN, so that it could control the system more efficiently.

The following variables were investigated to find the best operating point for the ANN:

- 1) ANN Learning rate (B),
- 2) ANN Voltage constant (Cv),
- 3) PWM switching frequency (Than),
- 4) ANN sampling frequency (Ts),
- 5) Alpha/Beta current and voltage feedback terms.

The results from these simulations showed that the desirable operating conditions were as follows:

ANN learning rate: $B = 0.12$
 ANN Voltage constant: $Cv = 1.1000323$
 Hanning sampling freq.: $Than = 10 \text{ kHz}$
 ANN sampling freq.: $Ts = 10 \text{ kHz}$

The results obtained from the optimal operating conditions are shown in Fig. 9 and Fig. 10 below.

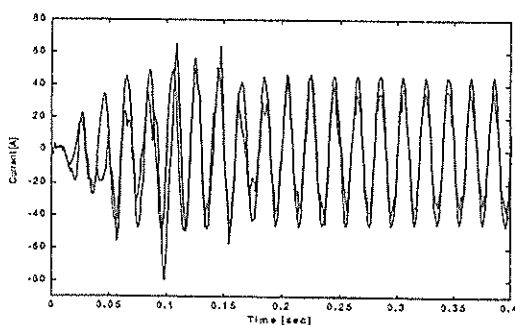


Fig. 9. Phase A supply current and ANN alpha current

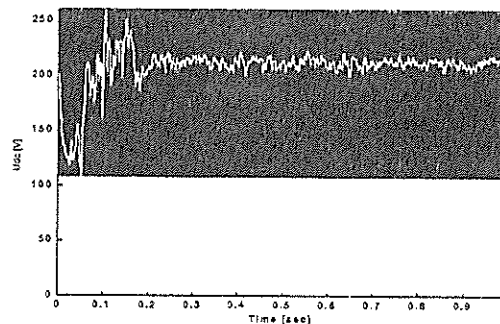


Fig. 10 DC Link Voltage

At this optimal operating point, the Boost rectifier retains a constant DC link voltage, which is required in VSD's, in order compensate for limited voltage dips on the mains supply. The simulation results therefore indicate that theory, a COT ANN current controlled boost rectifier can be utilized for voltage dip compensation in variable speed drives.

Thus for a COT ANN current controller to be able to control a Boost rectifier, it is imperative that it be finely tuned for efficient control, if this is not performed, the ANN will become unstable.

7. IMPLEMENTATION OF A REAL-TIME COT ANN CURRENT CONTROLLED BOOST RECTIFIER

For the practical implementation and verification of the design of the COT ANN current controlled Boost rectifier, a control platform was chosen that could realise the necessary control strategies. The ADC64 Digital Signal Processing (DSP) card [7] and Hyperceptions Hypersignal real-time integrated design environment [8] were chosen as the controller hardware and software platforms respectively, as they allow for fast data acquisition and control while using graphical representation of the control algorithm in the Hypersignal workspace. The DSP based current controller outputs alpha and beta controlling voltages, in the same manner as shown in Fig. 13, to a PWM card [9], which generates the necessary PWM switching pulses for the BOOST rectifier, to achieve the desired DC output voltage.

Fig. 12 shows the block diagram of the real-time COT ANN current controlled BOOST rectifier system used for practical verification of its viability as a voltage dip compensator. The outcome of the investigation of the real-time controller is discussed in section 8 below.

8 PRACTICAL RESULTS

Performance tests were performed on the real-time COT ANN current controlled Boost rectifier, in order to verify its effectiveness in achieving the objective of a constant DC output voltage, while maintaining sinusoidal line currents at unity power factor under fault conditions.

The variables and constants used for the practical system verification are identical to those derived for the CASED

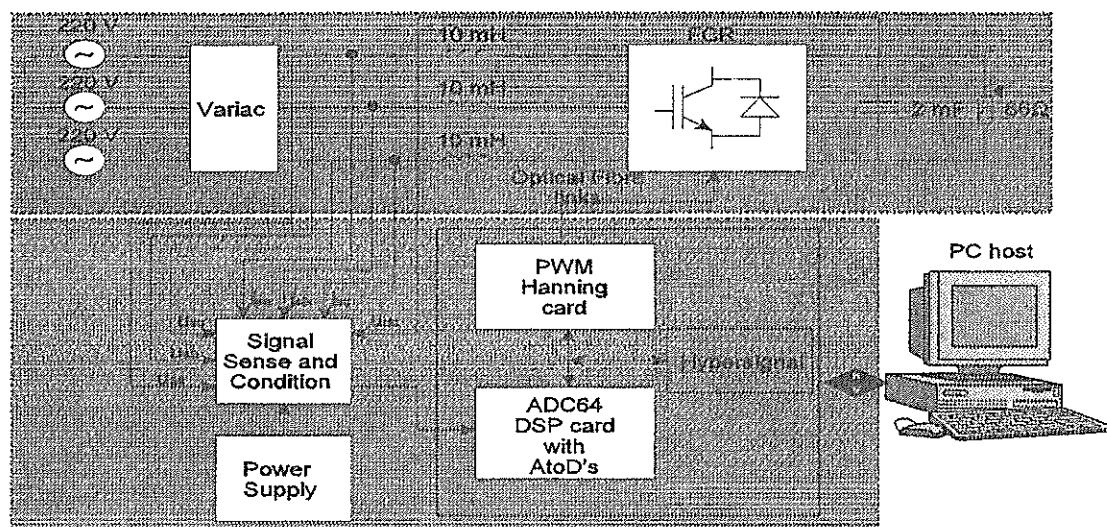


Fig. 12 Structure of practical system

simulations, shown in section 6. The results obtained under these operating conditions are shown in Fig. 11, Fig. 13, and Fig. 14, where the DC output voltage is being regulated to 167 V.

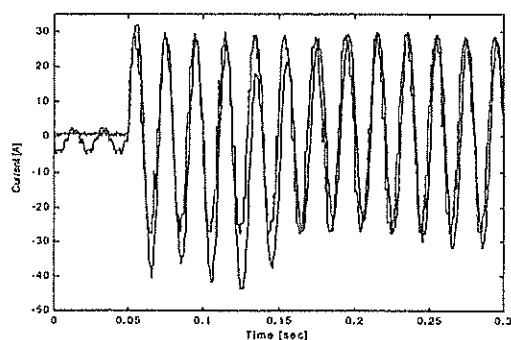


Fig. 11 Phase A supply current and ANN alpha current

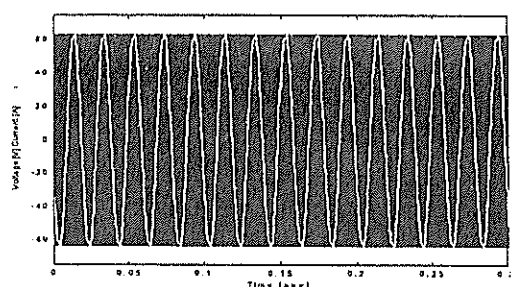


Fig. 13 Phase A line current and phase voltage

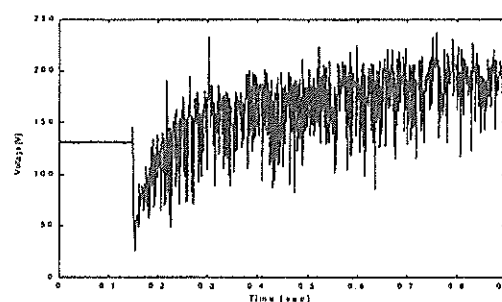


Fig. 14 DC Link voltage

The practical results show that the input phase A current converges to the ANN current controllers alpha reference current, and that it is sinusoidal and in phase with the phase A phase voltage. It also shows that the DC output voltage is regulated to the required level, thus it can be concluded that the COT ANN current controller may be successfully applied to a Boost Rectifier to reduce the effect of mains voltage dips. The oscillatory nature seen in the DC output voltage is due to limitations of the practical system at the time of publication.

9. CONCLUSION

Mathematical models were developed for a ANN current controlled Boost Rectifier, and simulated in CASE2 in order to investigate its viability, to overcome voltage dips. The practical and simulated results indicate that a COT ANN may be used to control a boost rectifier to compensate mains voltage dips. Future work will be to test the ANN controlled Boost rectifier system under voltage dip conditions, and to establish its limit of compensation. Furthermore, the COT ANN current controller will be compared to conventional current controllers, to determine

the most efficient and cost effective method for current control.

10. NOMENCLATURE

U _{sa}	A-phase phase to neutral voltage (V)
U _{sb}	B-phase phase to neutral voltage (V)
U _{sc}	C-phase phase to neutral voltage (V)
U _{dc}	DC link voltage (V)
E _α main	Supply alpha voltage component (V)
E _β main	Supply beta voltage component (V)
U _{ua}	A phase converter voltage (V)
U _{ub}	B phase converter voltage (V)
U _{uc}	C phase converter voltage (V)
U _{sd}	Supply d-axis voltage component (V)
U _{sq}	Supply q-axis voltage component (V)
U _{ud}	Converter d-axis voltage component (V)
U _{uq}	Converter q-axis voltage component (V)
U _{sα}	Supply α-axis voltage component (V)
U _{sβ}	Supply β-axis voltage component (V)
U _{uα}	Converter α-axis voltage component (V)
U _{uβ}	Converter β-axis voltage component (V)
MagU _s	Supply voltage magnitude (V)
I _{sa}	A-phase supply current (A)
I _{sb}	B-phase supply current (A)
I _{sc}	C-phase supply current (A)
I _{dc}	DC link current (A)
I _α main	Supply alpha current component (A)
I _β main	Supply beta current component (A)
I _{α_ref}	ANN alpha reference current (A)
I _{β_ref}	ANN beta reference current (A)
I _{ds_ref}	d-axis reference current (A)
I _{qs_ref}	q-axis reference current (A)
i _{sa}	A-phase instantaneous current (A)
i _{sb}	B-phase instantaneous current (A)
i _{sc}	C-phase instantaneous current (A)
i _{sd}	d-axis instantaneous current component (A)
i _{sq}	q-axis instantaneous current component (A)
i _α	Alpha instantaneous current component (A)
i _β	Beta instantaneous current component (A)

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THE APPLICATION AND HARDWARE IMPLEMENTATION OF CONTINUALLY ONLINE TRAINED FEEDFORWARD NEURAL NETWORKS FOR FAST ADAPTIVE CONTROL AND PREDICTION IN POWER ELECTRONIC SYSTEMS

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Abstract: This paper shows how continually online trained (COT) feedforward neural networks (NNs) can be used to solve adaptive control and prediction problems in power electronic systems, using the particular applications of induction motor and rectifier control as examples. These examples are also used to explain the need for fast COT NN hardware to realise these solutions in practise. Three COT NN hardware paradigms are described. The high backpropagation weight precision requirement, and the consequent lack of NN ASICs suitable for COT are also explained. The alternative of parallel digital implementation is discussed, and it is shown that, while significant advances have been made, this option may be too expensive for most potential applications. Finally, the development and performance of low precision COT algorithms are described, and the potential advantages of these algorithms for low cost analogue and hybrid VLSI implementation are explained.

Keywords: Neural Networks, Adaptive Control, Induction Motors, Power Converters.

1 INTRODUCTION

The self learning capabilities of feedforward neural networks (NNs) have been successfully applied to pattern recognition, time series prediction, and system identification and control. All of these functions are crucial in the advanced control of power electronic systems, since they are generally difficult or impossible to model by conventional methods due to complex nonlinear characteristics and large scale.

Where the required models are *time invariant*, the correct NN weights can be determined by *offline training* and then programmed into a number of commercially available ASICs, and deployed in power electronic applications such as

- frequency and harmonic spectrum analysis [1],
- rms waveform measurement estimation [2],
- improved PWM strategies ([3],[4]),
- converter fault detection/prediction [5], and
- motor fault detection/prediction [6].

This offline training approach can also be used to implement conventional power electronic control system functions, such as

- converter switching control ([7],[8]), and
- observers and estimators ([9],[10])

in order to exploit the speed and low cost of the VLSI hardware, as well as generic NN advantages such as noise immunity and fault tolerance.

However, most power electronic systems are either *under modelled* or are fundamentally *time variant*, and thus require NNs with *continual online training* (COT) in order to adaptively track *implicitly* and *explicitly* time varying characteristics respectively. These issues are discussed in section 2, where it is shown how COT NNs can be used to solve adaptive control and prediction problems in power electronic systems using the particular applications of induction motor and rectifier control. It is shown that the adaptation speeds in the order of microseconds required in both of these example applications are typical of power electronic systems, and the need for high speed COT NN hardware to achieve these adaptation speeds in practise is

explained.

Section 2I describes the three COT NN hardware paradigms, and explains the high backpropagation weight precision requirements and the consequent lack of suitable COT NN ASICs.

The alternative of high precision parallel digital implementation is discussed in section 4. The parallelisation problem posed by the massive connectivity of NN structure and COT is considered, and two methods of distributing the computational load with minimal communication overheads are described.

Finally, section 5 describes the development and performance of low precision COT algorithms, and explains the potential advantages of these algorithms for low cost analogue VLSI implementation.

2. COT NN APPLICATIONS IN POWER ELECTRONIC SYSTEMS

Adaptive control and prediction feedforward NN applications are based on their ability to identify dynamic system functions and track time variations in these functions by means of COT. Several methods have been developed to exploit these adaptive nonlinear modeling capabilities of NN system identifiers for:

- identifying inverse system dynamics and applying the trained NN model to achieve *direct inverse control* [11],
- *direct adaptive control* of systems where convergent training of NN controllers can be achieved by direct use of the output tracking error [12],
- *indirect adaptive control* by indirectly *tuning controller parameters* using NN identified system models [13] to map output tracking errors into equivalent control signal errors (where the control signal is the output of a NN, tuning of the controller parameters amounts to training of the weights of the NN controller to infer adaptive nonlinear control strategies [14]), and
- identifying suitable system models for *predictive control*

strategies [15], [16].

In power electronic systems, control action must be effected as fast as once in every switching period (usually less than 100 μ s) and take into account system variations which take place from one switching period to the next. NNs may be required to control not only total system dynamics, but also the power electronic devices, despite known and unknown nonlinearities and time variations in the system. Typical nonlinearities include transport lag, discrete switching and nonlinear mechanical and electrical loads. Time varying effects include parameter variations due to undermodeling of nonlinearities (certain transducer types are often omitted from control systems due to economic or technical limitations) and unpredictable load disturbances. All of these issues arise in the following two representative examples:

- Nonlinear adaptive control in induction motor drives, which falls into the above categories of *direct* and *indirect adaptive control*
- Time series prediction for improved performance of a current regulator for AC/DC converters, which falls into the above category of *predictive control*

2.1 Nonlinear Adaptive NN Induction Motor Control

The induction motor is an example of a complex, multivariable, nonlinear, dynamic electromechanical system which is rapidly time varying and thus requires fast adaptive control. While the scale of this problem is smaller than many control problems in the power industry, it is a good representation of the types of problems which remain to be solved.

The overall speed control objective is to control the flow of power into the motor so as to produce a torque on the rotor shaft which will result in a desired shaft speed response. This objective is usually realised by means of two controllers; a speed controller calculates the stator currents required to flow in the induction motor in order to produce the desired torque, and a current controller calculates the voltages which must be applied to the stator windings of the induction motor in order for the desired currents to flow.

The relationship between the applied stator voltage vector and the resultant stator current vector is *nonlinear* with respect to the speed of the motor shaft; this relationship is also *implicitly time varying* due to electrical parameter variations. In particular, inductances vary as a function of the stator currents due to magnetic saturation nonlinearities (inductances can thus vary significantly and rapidly due to significant and rapid changes in current magnitudes); resistances vary due to the skin effect (the AC resistance of conductors is frequency dependent and can thus vary rapidly with rapid changes in current frequency), and heating (changes in resistance due to heating effects are determined by thermal time constants and are thus generally relatively slow). In addition to the difficulties associated with current control, the relationship between the stator currents flowing in the motor and the mechanical torque produced by these currents is highly nonlinear and is also affected by time varying parameters and operating conditions. Furthermore, the shaft speed of the motor is determined by the response of the mechanical load (which may also be nonlinear and time varying) to the net torque acting upon it (the sum of the electromagnetic torque produced by the motor and any external disturbance torques). The dynamic performance of any position, speed or current control scheme therefore depends either on its robustness or adaptation to parameter

variations and the validity of any linearised models assumed in the controller design.

While robust control methods are specifically aimed at minimising the effects of model and parameter uncertainties, these uncertainties are generally too large or too system specific in the case of induction motor drives to obtain comparable implementation cost and performance to those achievable with adaptive control methods. Adaptive control, however, generally requires some form of system identification.

The stator current controller of [17] is based on continual online training (COT) of a NN to adaptively identify the NARMAX (Nonlinear AutoRegressive Moving Average with eXogenous inputs) model of the electrical dynamics of the induction motor. This model, derived in [17], gives the value of the next stator current sample $i_s(k+1)$ as some function, $f(\cdot)$, of the present "state" of the machine, $x(k)$, and the present voltage, $v_s(k)$, applied to the stator, i.e.

$$i_s(k+1) = f(x(k)) + c v_s(k) \quad (1)$$

where c is called the voltage constant and the state vector, $x(k)$, is a vector made up of present and previous values of stator current and shaft speed, ω_r , and a previous value of v_s , i.e.,

$$x(k) = [i_s(k), i_s(k-1), \omega_r(k), \omega_r(k-1), v_s(k-1)] \quad (2)$$

Equ. (1) shows that there exists some voltage, $v_s^*(k)$, which can be applied to the stator of the motor to produce some desired value of stator current, $i_s^*(k+1)$, i.e.

$$i_s^*(k+1) = f(x(k)) + c v_s^*(k) \quad (3)$$

Equ. (3) can thus be rearranged to produce a nonlinear control law:

$$v_s^*(k) = (i_s^*(k+1) - f(x(k))) / c \quad (4)$$

Thus the exact voltage required to produce the desired current can be calculated if the value of $f(x(k))$ and c are known. However, c and $f(\cdot)$ depend on the electrical parameters of the motor. Since these parameters are time varying, both c and $f(\cdot)$ are also time varying. Thus continual online training *must* be used in order for a NN to adaptively identify $f(\cdot)$ and c and produce approximate values $\hat{g}(\cdot)$ and $\hat{c}'$ respectively so that control equ. (4) can be implemented as

$$v_s^*(k) = (i_s^*(k+1) - \hat{g}(x(k))) / \hat{c}' \quad (5)$$

Equ. (5) is an example of *direct adaptive NN control* since COT is based *directly* on the control error, as shown in fig. 1(a). Fig. 1(b) illustrates the *indirect adaptive NN speed control* scheme of [17], where COT of the NN controller is *indirectly* based on the control error; an adaptive NN model of the current controlled induction motor is required to transform the shaft speed error into a corresponding error in the desired current commanded by the NN speed controller. The simulated current and speed control results of figs. 1(c) and (d) demonstrate the good adaptive tracking achievable with direct and indirect NN controllers respectively.

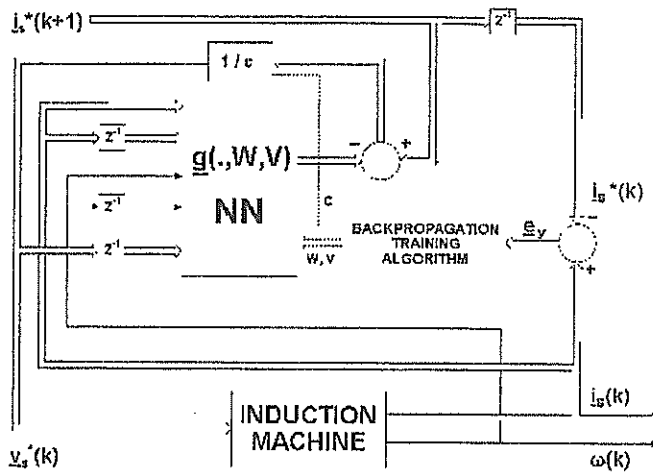


Fig. 1(a). Direct adaptive TM current control scheme

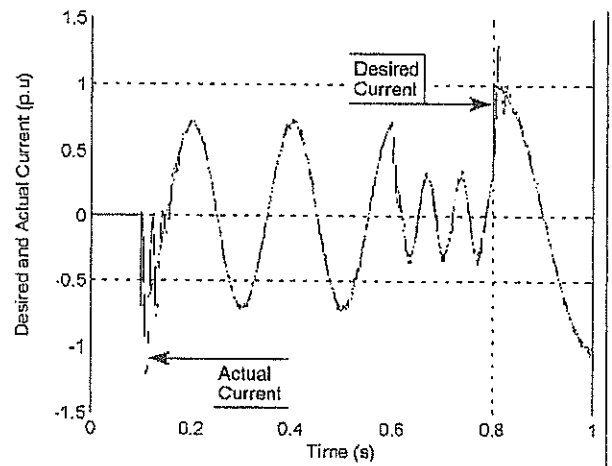


Fig. 1(c). Simulated NN IM current control performance

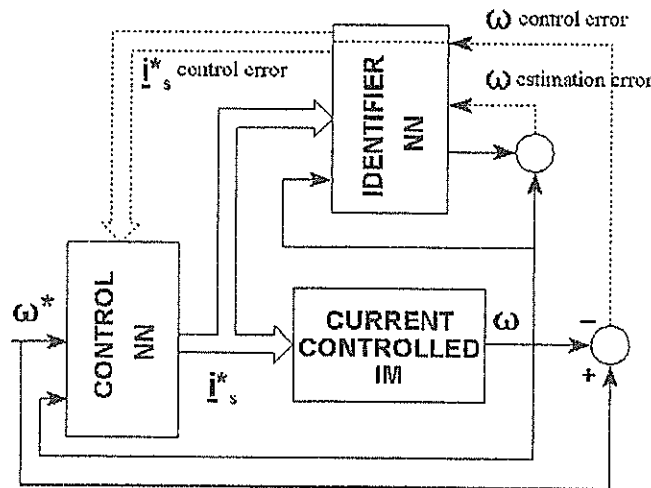


Fig. 1(b). Indirect adaptive TM speed control scheme

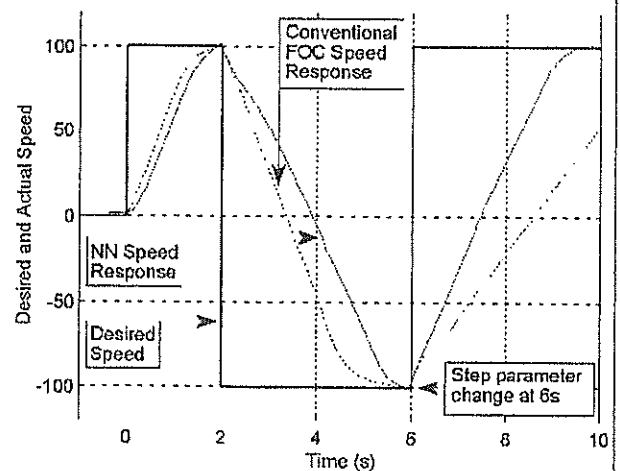


Fig. 1(d). Simulated NN IM speed control performance

2.2 NN Time Series Predictor Improved Rectifier Control

This section briefly describes an application of NNs in power electronics wherein the NN takes on a very different role compared to the induction motor control problem. This role represents the fastest type of forecasting problem encountered in power applications. Rectifiers (ac-to-dc converters) are an integral part of almost every power electronic system that connects to utilities which supply AC; this includes everything from small computer power supplies to transmissionline-connected FACTS elements. In the case of the three-phase bi-directional rectifiers, the boost-derived six-switch topology has become a popular choice. These rectifiers generally consist of two nested control loops; a slower outer loop that computes reference signals for the AC input currents and a faster inner current regulator that forces the actual AC input currents to follow these desired trajectories, as illustrated in fig. 2(a).

Predictive deadbeat control provides very good dynamic response with a fixed switching frequency [18], and can dramatically improve the performance particularly with small DC bus energy storage. The input current reference signals, $I_{dq}^{r,e}$ (fig. 2(a)), in the stationary dq frame of reference, are derived from a combination of the DC bus voltage error feedback and the output power

feedforward. This $I_{dq}^{r,e}$ is fed to a current regulator that uses the system model to compute the required voltage, v_{abc} at the rectifier side of the line inductors; v_{abc} is realised by space vector pulse width modulation in the rectifier. Because of the computation delay, by the time v_{abc} is calculated, other system variables i.e., the input currents, i_{abc} and voltages, v_{dq} , have changed to new values. In order to fully utilise the sampling time, T_s , for PWM purposes, this computed v is applied at the beginning of the following sampling interval. This results in a control error since the computation of v_{abc} was based on the actual (old) sampled system variables. However, if one-sampling-interval-ahead-predicted system variables are used in the control computation, it is possible to eliminate or reduce this error.

Another source of error stems from the fact that the $I_{dq}^{r,e}$ computation is based on instantaneous input power demand to supply the outward flow of energy and to compensate for any errors in the DC bus voltage. Because of the computation delay and the fact that the deadbeat control drives the input currents to the reference values in exactly one sampling interval, the input currents actually lag $I_{dq}^{r,e}$ by $4\pi T_s$, where T is the sampling interval. This lag results in the input power being slightly less than the desired power commanded by $I_{dq}^{r,e}$. If the DC bus energy storage capacity is small compared to the output power,

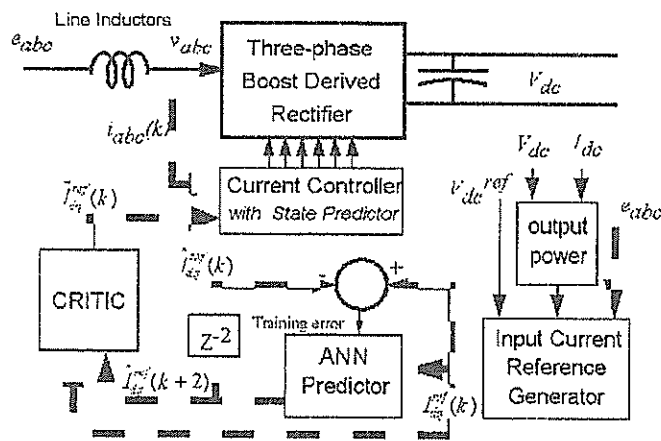


Fig. 2(a). Rectifier control scheme using a NN predictor

this delay manifests itself as small oscillations on the DC bus voltage. If the sampling time T_s is large or the DC bus capacitor is too small, these oscillations can cause instability. This error in deadbeat control can be cured by using a $2T_s$ ahead predicted value of I_{dq}^{ref} in the control computation. Because the I_{dq}^{ref} calculation requires the output power and the input voltages, which can not be modelled because of their statistical nature, this prediction of I_{dq}^{ref} must use statistical or AI methods. A NN based time series predictor can be used along with a state predictor [16] to greatly improve the performance of the rectifier regulator. Fig. 2(b) shows a typical result where the impact of the prediction schemes on the DC bus voltage ripple is obvious.

2.3 The Need for COT NN Hardware

The above two examples have illustrated two power electronic systems with time varying models, where the time variation is not explicit in the training data. These examples also show that time variations can be slow or very fast.

In the current loop of the induction motor control example, implicit time variations occur due to undermodelling of the electrical nonlinearities, which manifests itself as very rapidly time varying inductances. Furthermore, undermodelling of the skin effect and thermal dynamics of the motor results in unpredictable variations in the winding resistances. While these time variations could be accounted for by using additional sensors, these are undesirable because they increase cost and reduce reliability. In the speed control loop time variations include implicit mechanical load parameter variations, and explicit (unpredictable) external torque disturbances.

In the rectifier controller example, unpredictable rapid disturbances occur both on the load side and the power supply side. Furthermore, the random transport lag in the rectifier system cannot be statically modelled.

These examples are thus representative of the types of time varying characteristics which occur in power electronics systems in general. These examples also demonstrate that continual online training of NNs is imperative if they are to be used for adaptive control in these systems. Continual online training of NNs not only solves the nonlinear and time varying issues in system

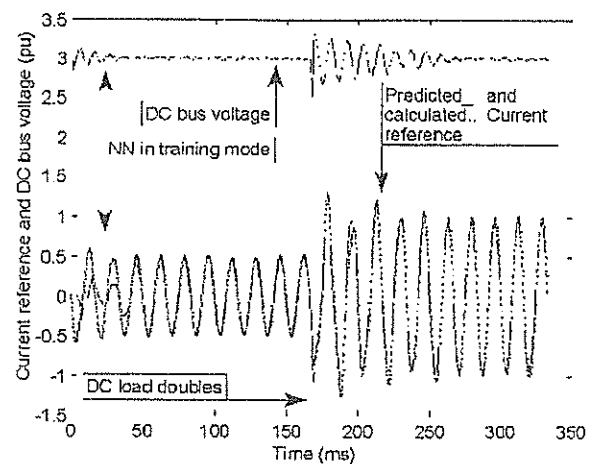


Fig. 2(b). Simulated regulator control performance

control, but also results in a controller which is totally *self-commissioning*. This is an important issue in many power electronic systems and machines. Tremendous value is added to the system if expert tuning of controller parameters is not required. Even conventional control systems with built-in adaptive tuning still require initial conditions for the parameters to be set, in order to have a satisfactory start-up. NN-based controllers, on the other hand, can start-up from random parameter values. The remainder of this paper deals with the hardware implementation of COT NNs in order to realise these controllers.

3 COT NN HARDWARE PARADIGMS

The most basic requirement for COT NN implementation is reprogrammable weights. The three main COT NN hardware paradigms are

- VLSI NN ASICs with off-chip DSP training,
- Single or parallel DSP implementations, and
- VLSI NN ASICs with on-chip training.

3.1 NN ASICs With Off-chip DSP Training

Most commercial and experimental NN ASIC hardware is now reprogrammable. ASICs such as the 80170NX [19] and WATTLE [20] can thus, in principle, be used for COT by implementing the training algorithm separately on a microprocessor and continually downloading new weights to the trainable ASIC. While this is not the originally intended normal mode of operation of trainable ASICs, the basic mechanism just described has actually already been implemented by [21] (using the 80170NX), and several others [22], to perform so called *chip in the loop* (CIL) training. CIL training is essentially an offline training method employed to specifically tailor the training process for the exact characteristics of a particular type of ASIC, which may be difficult or time consuming to simulate and could vary significantly from one manufactured device to the next of the same particular type. COT, however, is a fairly specialised NN paradigm, which requires that the training be done in a specified amount of time. CIL hardware may not be fast enough to achieve the speed of COT required for some power electronic applications due to large NN size and/or rapid time variations. The main reason for the limited speed of COT achievable with CIL

hardware is the weight download process; this process could entail simple DSP routines for digital ASICs, or more complicated procedures for analogue or mixed signal devices. In order to avoid the weight download bottleneck of CIL hardware, either the NN feedforward function must be moved onto the DSP, or the training algorithm moved onto the VLSI ASIC.

3.2 Single and Parallel DSP COT NN Implementations

DSPs are ideal for fast vector processing, and are thus well matched to the computational requirements of COT NNs [23]. DSP implementations are also very flexible, and allow incremental changes to working installations, and fundamental research and development work, such as the practical evaluation of different NN architectures and training algorithms. However, some applications may require COT rates which approach the limits of single high power DSPs, or may simply be more economically viable if several cheaper DSPs can be used in parallel.

3.3 NN ASICs With On-chip Training

There is a growing trend in VLSI feedforward NN ASIC technology towards on-chip training. [24] and [22] describe devices with on-chip training which have been implemented in digital, analogue, and hybrid (mixed signal) VLSI NN circuits.

Digital VLSI: The relatively low density of digital VLSI circuitry places severe restrictions on the size of NN which can be implemented on a single device. Hitachi [25] addressed this problem using wafer-scale integration to achieve a NN size of 144 neurons, with a backpropagation weight update speed of 300 M weights per second. However, as with all digital VLSI NN ASICs, NN size and speed must be traded off against the bit precision, which is 8 bits for the weights, and 9 bits for the activation functions in this device. The activation function is ideally bounded between +1 and -1; its precision is thus not as significant as that of the weights, which are ideally unbounded. It has been shown [26] that 22 bit weight precision is required for backpropagation training in general. Digital ASICs with on-chip backpropagation training are thus generally suitable only for offline training applications, where convergence may be induced by trial and error settings of various training parameters. Limited precision NNs are generally also suitable only for inexact applications such as pattern recognition, and may be inadequate for applications where nonlinear functions must be accurately learned and approximated. However, some digital VLSI NN devices with on-chip training can be connected in parallel to implement NNs of larger size and/or higher precision.

Analogue VLSI: Analogue NN ASICs exploit the physical electrical properties of materials to implement NN operations with very high speed and VLSI density. [27] reports on an analog on-chip implementation of the backpropagation training algorithm. The effective weight precision of analogue VLSI, however, is typically only as high as 6 bits [22] due to parasitic effects such as charge leakage, offsets, and thermal variations. Furthermore, the characteristics of individual weight storage, activation function, and multiplier implementations can vary significantly on the same chip due to fabrication difficulties. Analogue on-chip backpropagation is thus also generally suitable only for offline training pattern recognition applications.

Hybrid VLSI: Hybrid designs attempt to combine the external interfacing and weight precision and storage advantages of digital circuitry with the speed and density of analogue VLSI. [28] reports on an analogue NN ASIC with EEPROM weights, and an on-chip implementation of a new algorithm called random weight change (RWC) training. RWC is a *stochastic* gradient descent algorithm specifically designed to overcome the convergence problems of *exact* gradient descent algorithms (such as backpropagation) associated with the limited weight precision of analogue circuits. Another hybrid approach involves the use of various pulse stream modulation techniques. The Ricoh Company have produced [29] a pulse stream NN ASIC with on-chip backpropagation training. However, while pulse stream devices generally have higher effective precisions than other hybrids and pure analogue devices, the high weight precision of 22 bits required for stable COT with the backpropagation algorithm cannot be achieved with this technology. The two COT NN paradigms suitable for fast COT are therefore high precision parallel digital implementations, and NN ASICs with on-chip training algorithms such as RWC, which do not require high precision weights for stable convergence.

4 HIGH PRECISION PARALLEL DIGITAL COT NN IMPLEMENTATION

Several digital VLSI NN architectures facilitate the digital hardware implementation of NNs with on-chip training of larger size and/or higher precision by connecting building blocks in parallel:

- 8 Hitachi wafer-scale digital VLSI NN ASICs has been used [24] to construct a NN with 1152 neurons and a backpropagation 16 bit weight update speed of 2.3 G weights per second, and
- single chip SIMD (single instruction multiple data) processor arrays can be used to build NNs with 64 neurons and backpropagation 16 bit weight update speeds of 300 M weights per second [30].

While relatively large NNs can be realised in this way, the weight precision appears to be limited to 16 bits, whereas 32 bit or higher precision can be easily achieved using standard DSPs in parallel.

Fig. 3(a) depicts a typical mapping of the feedforward NN structure onto standard multi processor hardware. It can be seen that many connections are cut due to the highly connected NN structure. It has been shown [31] that the large number of resulting communication transactions place an upper limit on the gain in speed at 1.7 times for 5 or more processors. Figs. 3(b) and 3(c) depict two methods which have thus been developed to increase NN parallelism for fast COT [23].

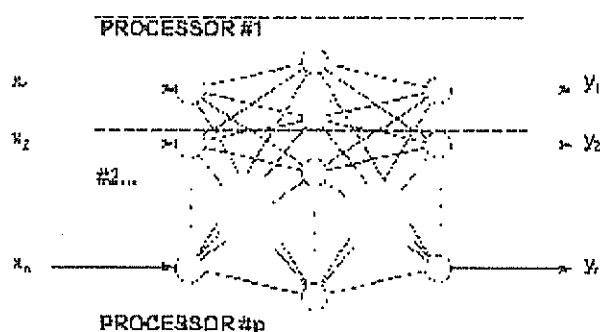


Fig. 3(a) A typical standard multiprocessor NN mapping

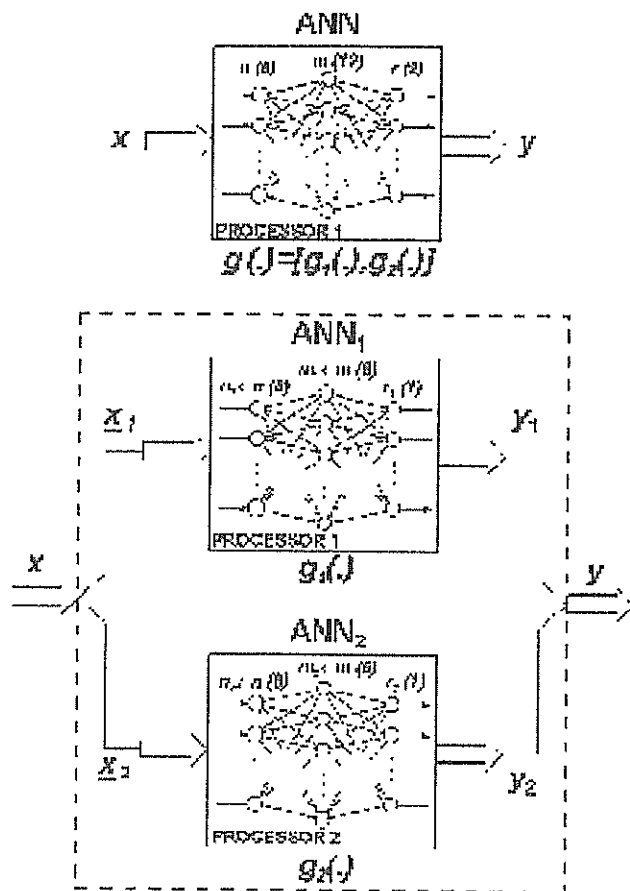


Fig. 3(b) A pair of separated output NNs

The first method is called *output separation* since a single multiple output NN is replaced with *separate* smaller NNs which each calculate one output. Each of these NNs are then implemented on separate processors which need not communicate with any of the others. In the case of the NN current loop, the vector valued input/output mapping function $g(\cdot)$ of equ. (2) has two components $g_1(\cdot)$ and $g_2(\cdot)$. The speed of execution is *doubled* by replacing the single 12 hidden neuron NN used to simultaneously identify both functions, with two separate 6 hidden neuron NNs, as shown in fig. 3(b). The second method exploits the fact that the feedforward or output calculation and backpropagation training parts of the NN operation take approximately equal lengths of time when implemented on DSP hardware. It was also found that control performance did not degrade significantly if only one training cycle was done for every two forward or output calculation cycles. Fig. 3(c) shows how the standard (original) speed of execution can be *doubled* by exploiting this fact. Here two identical NNs are implemented on separate processors, called a *tandem pair* since they operate in tandem. One NN does the forward or output calculation while the other trains and visa versa. Once again, each NN operates independently of the other and there is no need for communication between them. Each executes its forward or output calculation based on training information from every second cycle to effectively achieve the desired ratio of one training cycle for every two forward or output calculations. [32] presents the output separation and tandem COT parallelisation methods in more detail, and shows practical results which verify that neither method significantly alters the overall NN

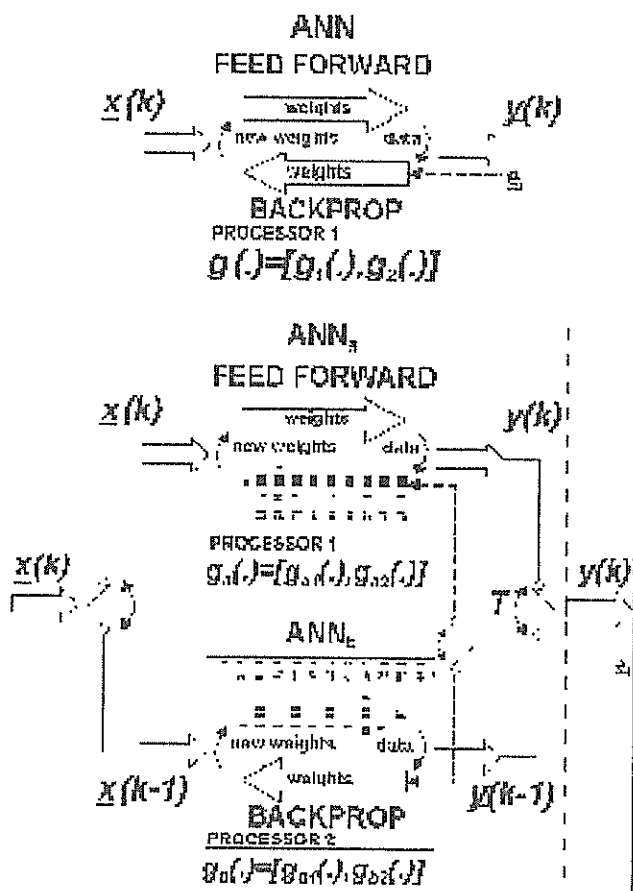


Fig. 3(c) A tandem pair of NNs

convergence or induction motor current control performance.

Furthermore, the output separation and tandem parallelisation techniques can be combined to achieve an optimal linear speedup of four times with 4 DSPs for the 8x12x2 induction motor control NN. In general, these methods can be combined to yield speedup factors approximately equal to twice the number of NN outputs; NNs with tens or hundreds of outputs (and correspondingly larger hidden layers and input vectors) could thus be parallelised to achieve speeds as much as one or two orders of magnitude faster than single DSP implementations.

However, the tens or hundreds of DSPs required to achieve these speeds of COT may be economically justifiable only as a development cost in many applications. Such applications would thus benefit from NN training algorithms suitable for digital, analogue, or hybrid VLSI on-chip implementation with limited precision.

5 ALGORITHMS FOR FAST ON-CHIP COT WITH LIMITED PRECISION

Several *approximate gradient descent* and *stochastic gradient descent* algorithms have been developed to overcome the precision limitations of analog VLSI hardware and some of these have been successfully implemented for on-chip training of mixed

signal NN ASICS.

MR III (MADALINE RULE III) [33] is an example of an approximate gradient descent algorithm. This rule basically calculates the change in a given weight based on an approximate "measure" of the gradient of the global error function; the gradient is approximated by the observed change in the error function which results from a *perturbation* of the *activation* of a given node. MR III is thus an improvement over backpropagation, but in large NNs the sequential gradient measurements may be too time consuming for fast COT.

Stochastic gradient descent methods such as Random Weight Change (RWC) [28], ALOPEX [34] and SWP [35] do not require exact knowledge of the gradient of the output error function. These methods find the approximate gradient from a random search, instead of computing or measuring its value. ALOPEX (essentially a batch training method) searches for the gradient on the global error surface, and requires more computations than RWC (a delta rule method), which searches for a gradient on a local error surface. In its original form, however, RWC is suitable only for CIL training but not for COT. A COT form of the RWC algorithm has thus been developed, and is described below.

5.1 The COT RWC Algorithm

The convergence mechanism of COT RWC is best explained by comparison with that of backpropagation training. The backpropagation training algorithm is merely a set of equations which are a closed form solution to the *gradient descent* rule as applied to each of the input and output weights of the NN, i.e.

$$\Delta w_{ij} = -k \frac{\partial E(\bullet)}{\partial w_{ij}} \quad (6)$$

where $k > 0$ is known as the *learning rate* (and is heuristically determined for optimal convergence) and $E(\bullet)$ is the scalar function:

$$E(\bullet) = \frac{1}{2} \| \mathbf{f}(\mathbf{x}) - \mathbf{g}(\mathbf{x}, \mathbf{W}, V) \|^2 \quad (7)$$

In other words, the *backpropagation algorithm* calculates the *exact gradient* of $E(\bullet)$ with respect to each of the weights and changes them in the *negative* direction of this gradient so as to reduce the value of $E(\bullet)$, which corresponds to improving the approximation of the desired function $\mathbf{f}(\bullet)$ by the NN function $\mathbf{g}(\bullet)$.

Instead of *calculating* the exact gradient of $E(\bullet)$, the COT-RWC algorithm *randomly searches* for the gradient in the following way:

1. The value of $E(\bullet)$ is stored as E_{best} and the stored matrices Δw_{best} is initialised to zero.
2. Each of the NN weights are changed by a fixed amount, i.e.

$$\Delta w_{ij} = \rho \delta \quad (8)$$

where δ is some *small positive* constant (roughly analogous to the learning rate in backpropagation training) which is

heuristically determined for optimal convergence, and ρ is a random variable which may assume either the value of +1 or -1 with equal probability.

3. The new value of $E(\bullet)$ is compared with E_{best} . If $E(\bullet) < E_{best}$, $E(\bullet)$ is stored as E_{best} and all of the weight changes are stored in Δw_{best} . (CIL-RWC simply changes the weights if this condition is true and continues to add this change until it results in an increase in error).
4. Each of the NN weights are restored to their original values (the values they possessed at step 1).
5. Steps 2, 3 and 4 (collectively known as one random trial) are repeated $N-1$ times.
6. Each of the NN weights are changed by the corresponding values stored in matrices Δw_{best} .

The total *number of trials* N is determined beforehand by the time T within which each training cycle (one evaluation of steps 1 to 6) is required to be completed. For adaptive identification and control of induction motor currents, T corresponds to the portion of each sampling period which may be devoted to training. It is expected [36] that particular mixed signal VLSI implementation of COT-RWC will allow $N=20$ for $T=50 \mu s$, which will correspond to a sampling rate of approximately 10 kHz. Under these circumstances, in each training cycle, the COT-RWC algorithm will choose a direction in which to change the NN weights from a set of twenty different random directions. The direction which is chosen will be that which will produce the greatest decrease in the value of $E(\bullet)$. Thus the direction which is chosen will be that which is *closest* to the exact negative gradient of $E(\bullet)$.

Since backpropagation always changes the weights in the exact direction of the negative gradient (the steepest descent) it is reasonable to expect that backpropagation may converge faster than RWC. Nevertheless, COT-RWC will eventually converge since there is always a finite probability that a direction will be found in which the weights can be changed so as to reduce the value of $E(\bullet)$. The question of whether or not the convergence of COT-RWC is sufficient to produce rapid and reliable control of power electronic systems has been investigated by means of the induction motor and rectifier control application examples. [36] and [37] present results which show that COT RWC produces very similar performance to backpropagation training for both induction motor current control, and rectifier control respectively.

5.2 RWC vs Other Algorithms for COT NN ASICS

ALOPEX [34] is similar to RWC in that $dw_{ij}(k) = +\delta_{ij}(k)$ with probability P_{ij} and $dw_{ij}(k) = -\delta_{ij}(k)$ with probability $1-P_{ij}$. However, RWC uses a constant $P_{ij} = 1/2$ whereas ALOPEX calculates

$$P_{ij} = 1/[1 + \exp(-C_{ij}/T)], \quad (9)$$

where

$$C_{ij} = [w_{ij}(k-1) - w_{ij}(k-2)][E(k-1) - E(k-2)] \quad (10)$$

is the correlation between the previous change in weight and the previous change in error (cost function value) and T is a positive temperature. The parameter T , thus effectively determines the difference between COT-RWC and COT-ALOPEX. For large T , $P_{ij} = 1/2$. As T decreases, so does the probability increase that

ALOPEX weight changes will be either in the same direction as previous weight changes if these weight changes reduced the error, or in the opposite direction to previous weight changes if these weight changes increased the error. That is, the weight changes become less random and more related to the previous change in error with decreased temperature. The convergence of both RWC and ALOPEX is dependent on the step size, δ . The apparent advantage of ALOPEX over RWC is the possibility of improved convergence due to explicit biasing of the random weight changes in the direction of decreasing error. However, the present form of ALOPEX is suitable only for CIL since it does not employ a direct mechanism to ensure that weights are actually never changed either in the direction of increasing error, or by too large a step size in the direction of decreasing error so as to overshoot minima and actually produce an increase in error. While the technique employed in COT RWC to overcome this problem (only actually applying the best of N purely random $P = 1/2$ weight changes) could be adapted and used to produce a COT ALOPEX algorithm, this same technique inherently provides COT RWC with a crude approximation of the weight change biasing mechanism employed by ALOPEX. Furthermore, the ALOPEX biasing mechanism is more complex (and may not be viable) for on-chip mixed signal implementation in its present form. However, RWC may be improved by including an explicit approximation of the ALOPEX correlation effect of equ. (10). This could be achieved by reevaluating the error function $E(\bullet)$ for the reverse of random trial which produces an increase in $E(\bullet)$.

SWP (summed weight perturbation) [35] has been implemented on an analogue ASIC [38]. The algorithm is similar to RWC in so much as the approximate gradient is recursively searched for by means of random weight perturbation. However, SWP is more of a stochastic version of MR III because each of the input weights associated with a particular neuron is perturbed by the same random amount. The net effect is a random perturbation in the activation of each neuron. As such the resolution of the gradient search is significantly less than that of RWC. While the SWP gradient search may thus be faster to implement, and hence allow more random trials in each epoch, the limited resolution may result in significantly lower convergence than RWC. Hardware implementations of SWP may therefore be less suitable for COT than RWC devices.

5.3 Hardware Implementation

Limited precision COT algorithms such as RWC can be implemented on any hardware with reprogrammable weights in order to overcome the backpropagation convergence problem. However, the speed of implementation (in terms of weight updates per second) will be higher for analogue and hybrid ASICs than for Digital ASICs. The implementation cost will also be lower because the existing analogue weight refresh circuits in these devices can easily be modified to implement weight perturbation. Promising candidate devices include

- the analogue TARANA ASIC of [35],
- the hybrid CIL RWC ASIC of [28], and
- the pulse stream hybrid ASIC of [39].

6. CONCLUSION

This paper has considered the particular applications of induction motor and rectifier control to demonstrate the complexity, nonlinearity, and time varying characteristics typical of models required for good adaptive control in power electronic systems. It also has used these example applications to show how neural networks (NNs) can identify these models, and explained the need for continual online training (COT) to track time variations. Simulation results have been presented to demonstrate the potential self-commissioning and good adaptive control performance achievable with COT NNs. Three COT NN hardware paradigms have been described, and it has been shown that on-chip implementation of limited precision training algorithms can be achieved through minor adaptations to existing hybrid VLSI devices to produce low cost NN ASICs capable of fast COT.

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Determination of Loss Load Factors in the Western Cape Network

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Abstract: It is estimated that I^2R losses cost South Africa of the order of R2 billion a year. Eskom has initiated a programme to determine where these losses are occurring, and what can be done to reduce the extent of the problem. A key factor in this study is the determination of loss load factors for various system components (so as to be able to derive average loss estimates from peak loss figures). This paper deals with initial work in determining loss load factors, using Eskom's PSS/E (Power Systems Simulator for Engineers) load flow package to study the Western Cape network.

Keywords: Loss Load Factor, Distribution Network, Main Transmission Substation

1. INTRODUCTION

It is a well-established fact in any power system that there is a discrepancy between the energy sent out of the generation plants and the energy consumed. The discrepancy, which is termed as loss, can be attributed to technical and non-technical reasons. Theft, meter reading and metering errors account for non-technical losses. The technical losses can be broadly categorized into 'fixed' and 'variable' losses. The fixed losses such as transformer iron losses and dielectric losses of cables are mainly voltage dependent. The fixed losses are very small relative to the variable, and are not considered further in this work. The variable losses are due to current flowing in the resistive components of the plants. Since current varies with the load the losses vary accordingly.

This thesis project is concerned with technical losses in the Distribution System of the Western Cape. The voltage levels currently being used in the Western Cape system are 400kV, 132kV, 66kV, 33kV, 22kV, 11kV, 6.6kV and 3.3kV. Although there are very few consumers being supplied at transmission HV voltage of 400kV, it is normally understood that distribution is mainly concerned with systems outages including and below 132kV.

This paper is based on work done by the first author as part of a final year thesis project in 1998.

2. PROBLEM DESCRIPTION AND FORMULATION

The loss load factor (LLF) is defined as the Average Losses divided by the Peak Losses. The load factor (LF) is defined as the Average Demand divided by the Maximum Demand. The loss load factor (LLF) is related to the load factor (LF) by means of the following equation:

$$LLF = \alpha * LF + \beta * LF^2$$

where

α = peak load constant

β = off – peak load constant

$$\alpha + \beta = 1$$

If one considers the energy consumption of a network for a certain period, e.g. one month, and a given load factor (LF), two extreme cases can be worked out. In the first instance peak power is consumed for a fraction of the total time period. Since **Power Loss** = I^2R , the greatest amount of energy will be lost in this scenario [1].

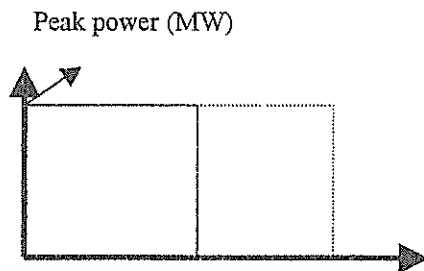


Figure 2 A: showing one extreme case where $LLF = LF$

Thus I refers to the peak current in the equation **power loss** = I^2R , and it can be said that the **Peak Power Loss (Ppl)** = I^2R , and the energy loss in this case be **Ppl*LF**. The losses are thus proportional to LF in this case [1].

In the second instance the load is consumed continuously at the fraction LF of peak power

(i.e. $LF*I$ is flowing) except for a short peak spike [1].

Peak power (MW)

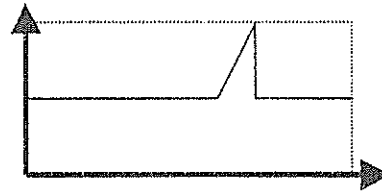


Figure 2 B: showing the second extreme case, where $LLF = LF^2$

This constitutes the best case with respect to losses. Since $LF*I$ is flowing,

$P_{pl} = (LF*I)^2R$. The losses are thus proportional to the LF^2 in this case.

For any load profile it is true to say that the losses will definitely be somewhere between the best and the worst case [1]. Hence we know $LF^2 \leq LLF \leq LF$.

2.1 The meaning of the constants α and β

The constant α is defined as the peak load duration and β the constant of off-peak load duration. From figure 2.1a the peak load duration $\alpha = \alpha_1 + \alpha_2$ and the off-peak load duration $\beta = \beta_1 + \beta_2 + \beta_3$.

It is even more accurate to calculate α and β by transforming the load curve into a load duration curve. Rearranging the load points into descending order produces the load duration curve. The sequence in which the load points occurred is lost but much is gained for the purpose of our analysis.

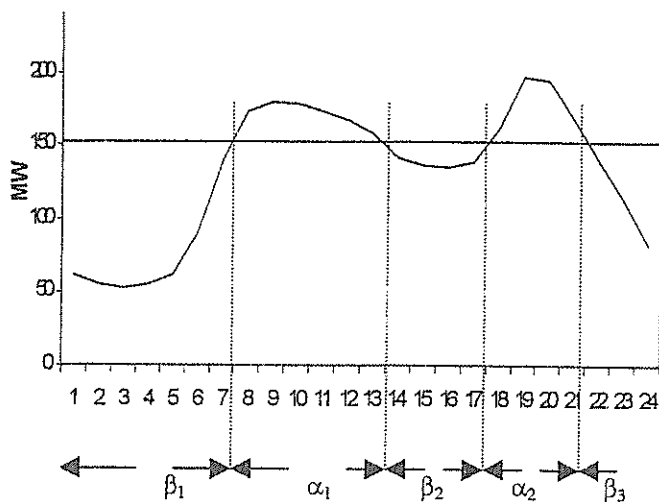


Figure 2.1a: typical daily load curve

Taking the Average between the Maximum Demand and the Average Demand produces a load point that will approximately divide the area of the curve into two equal areas. Since α and β have to be equally distributed through the area of the load curve, we conclude that α must occupy half the area of the curve (to the left since it is where peak load points occur) and β the remaining half. **N.B:** This does not mean α and β are equal.

2.2 Procedure for calculating Distribution Technical Losses

Refer to the Standard Draft for Calculating Losses in the Distribution Business [1].

The total load for a Main Transmission Substation (MTS) Network is taken as the measured load at the Point of Common Coupling (PCC) on the boundary between Transmission and Distribution. The hourly load readings are available from METMAS system at Transmission Economics or from the load Data Warehouse in the Electricity Demand Department (EDD) of ESKOM. system at Transmission Economics or from the

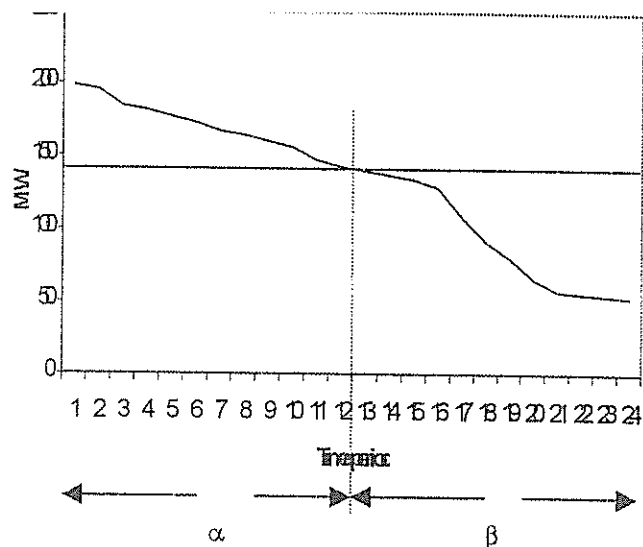


Figure 2.1b: typical daily load duration curve

load Data Warehouse in the Electricity Demand Department (EDD) of ESKOM.

Scaling of the Base Case (100% Reserved Capacity per MTS) load condition is done in proportionate to the actual MTS load for the Accounting Period (load cycle) in question e.g. If the Base Case load for a particular MTS is 100MW total for a group of 20 customers and the actual load 80MW, each one of the 20 individual loads need to be scaled down to 80% of their Base Case Values.

If the “scaled” load of an individual point of supply differs more than 20% from the known actual load for the same point, the actual known value is used instead.

The **Accounting Period** (the period for which the load data is available) that will be used for this analysis is a week (7 days).

3. CASE STUDY

The Distribution Network of the Western Cape where studied using the methods as described above. The network consists of 15 Main Transmission Substations (MTS). Four of the fifteen are in the Peninsula region, three in the West Coast region, three in the South Cape

region and the remaining in the Springbok region. The load flow studies were done using PSS/E (Power Systems Simulator for Engineers). The results of this study are tabulated in table 2.2.5 D.

	Substation Name	Distribution copper losses (line & transformer) (MW)
1	Acacia	355.15
2	Koeberg	187.44
3	Muldersvlei	2588.36
4	Bacchus	1719.06
5	Proteus	1410.75
6	Droenvier	233.23
7	Aurora	201.76
8	Helios	6.59
9	Juno	13.19
Total		6717.53

Table 2.2.5 A: Total Distribution copper losses for all Main Transmission Substations

Substation Name	Total transformer No - load losses (MVA)
Acacia	0.28658
Koeberg	0.1088
Muldersvlei	1.42405
Bacchus	0.57538
Proteus	0.28617
Droenvier	0.04955
Aurora	0.3317
Helios	0.05038
Juno	0.0858
Total	3.19841

Table 2.2.5 B: Total No – load losses of all Main Transmission Substations

4. CONCLUSIONS

The results show that Distribution Technical losses could be calculated using the methods proposed in this paper. In the authors' opinion these methods should have been adopted long ago, since it is not something that requires great technical expertise.

	Substation Name	Total transformer load losses (MVA)
1	Acacia	2596.94
2	Koeberg	1095.36
3	Muldersvlei	11018.45
4	Bacchus	3584.78
5	Proteus	2443.06
6	Droenvier	103.84
7	Aurora	1829.86
8	Helios	199.38
9	Juno	322.56
Total		23196.43

Table 2.2.5 C : Total Transformer load losses for all Main Transmission Substations

Substation Name	Sub ID	Reserved Capacity MW	Peak Load Losses MW
Acacia	376	566	2.41
Koeberg	392	141	1.28
Muldersvlei	391	862	17.59
Philippi	381	557	-0.09
Aurora	388	163	1.86
Helios	385	18	0.15
Juno	379	36	0.11
Bacchus	387	248	12.42
Droenvier	394	47	1.69
Proteus	382	283	9.51
TOTALS		2921	

Table 2.2.5 D: Peak power losses for 100% reserved capacity

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AN INVESTIGATION INTO THE TECHNOLOGICAL AND ECONOMIC ASPECTS OF COMPACT FLUORESCENT LAMPS

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Abstract: For a similar light output, compact fluorescent lamps consume about a fifth of the real power of incandescent lamps. However, electronically-ballasted CFLs produce a wide range of harmonics, which may cause voltage distortion on a supply network, or excess heating in distribution equipment such as transformers, conductors and circuit-breakers. Revenue meters are also susceptible to registry problems in the presence of CFL harmonics. From an economic perspective, the use of CFLs is very favourable to both Eskom and the customer. The environment can also benefit by the use of CFLs through a reduction in pollution and a conservation of natural resources.

Keywords: Compact fluorescent lamp, electronic ballast, harmonics, payback period.

1 INTRODUCTION

In South Africa, there is an ever-increasing demand for electricity. In particular, South Africa's commitment to the National Electrification Scheme, which forms a major part of the RDP (Reconstruction and Development Programme), has contributed largely to this growing demand. For example, South Africa's electrical demand increased by 11.3% to 27 967 MW in 1996 from 1995 [1].

One way of meeting this growing demand is for Eskom to increase its generating capacity by building more power stations. However, at present, installing one megawatt of generating capacity costs about five million Rand. Clearly, this method of catering for the electricity demand is costly, and can place a substantial burden upon the environment.

An alternate way of meeting this demand is through Demand-Side Management (DSM). From an end-user's point of view, DSM promotes the use of demand-reducing and energy-saving technologies, which may help eliminate or delay the construction of new power stations. Eskom's Residential Demand-Side Management (RDSM) is investigating several ways of reducing the demand of electricity. Since lighting contributes significantly to the national peak demand, the RSDM team is looking at the use of compact fluorescent lamps (CFLs), to replace incandescent lamps in households.

For a similar light output, CFLs consume much less real power than incandescent lamps. This offers potential savings to both Eskom and the end-user. However, the CFLs presently available on the South

African market produce a wide range of harmonics, which could lead to problems on the supply network.

The aim of this paper is to quantify the amount of harmonic distortion produced by CFLs, and to discuss the effects of these harmonics on various components of a distribution system. Also, an economic analysis is carried out to determine the extent of savings that can be incurred through the use of CFLs.

2 HARMONICS FROM CFLs

2.1 Electronic ballasts

All fluorescent lamps have a negative current-voltage characteristic. As a result of this characteristic, fluorescent lamps require a ballast to limit the current passing to the lamp. There are two main types of ballast used in CFLs:

- line frequency, electromagnetic ballasts; and
- high frequency, electronic ballasts.

The electromagnetic ballasts, also referred to core-and-coil ballasts, are usually simple combinations of a 50Hz autotransformer, and one or two inductances. Because these ballasts are operated at line frequency, they require a large inductance in the order of 1 Henry, to regulate the current to the lamp, as well as to supply the high voltage pulses necessary for lamp ignition. The disadvantages of using magnetic ballasts for CFLs are:

- the weight of the ballast due to the large inductances required; and
- the power losses that occur, mainly due to core losses in the 50Hz magnetics.

These losses consume up to 30% of the lamp's rated power, and severely reduce the efficiency of the lamp [2].

For these reasons, most manufacturers of CFLs opt for the high frequency, electronic ballast. By increasing the operating frequency from 50Hz, to say 50kHz, the ballast inductance is reduced by a factor of 1000 (1mH instead of 1H). Thus, by operating at a higher frequency, smaller inductances are used and the weight of the ballast is reduced. Furthermore, operating at a higher frequency leads to an increase in luminous efficacy.

Although the use of a high-frequency electronic ballast improves the efficiency of the lamp, the amount of harmonic distortion produced from the lamp increases as well.

In operation the electronic ballast is a frequency-converting, ac-to-ac converter. The ac power at line frequency is applied to an input rectifier circuit. The resulting dc voltage is applied to a high-frequency electronic switch that feeds chopped dc to a high-frequency power transformer. The transformer, in turn, delivers the required ac voltage to the lamp, at the switching frequency.

The front end of a basic electronic ballast is shown below in Figure 1. [3]

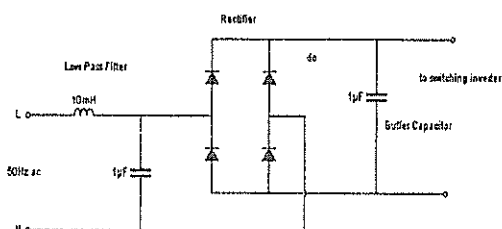


Figure 1: Input stage of an electronic ballast

The harmonics produced from CFLs result from the combination of the full-wave rectifier bridge and the buffer capacitor, in the input stage of the electronic ballast. The bridge conducts during the period of the cycle when the line voltage exceeds the capacitor voltage. The effect is that current is drawn from the power source over a very short time, as shown in Figure 2. Since the line current is no longer sinusoidal, it results in the generation of harmonics.

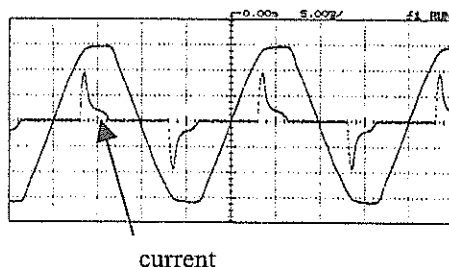


Figure 2: Voltage and current waveforms for an electronic ballast

A typical harmonic amplitude spectrum is shown in Figure 3 below.

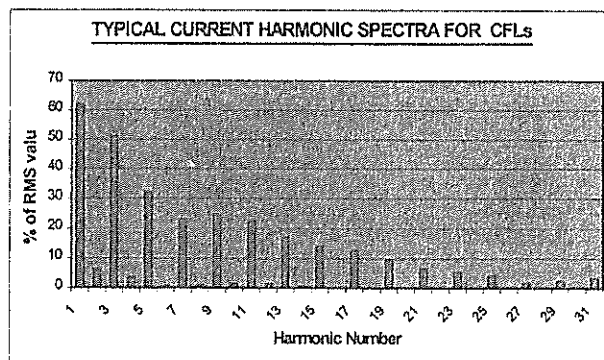


Figure 3: Harmonic spectrum for electronically-ballasted CFLs

As can be seen, the harmonic spectrum constitutes of mainly odd harmonics. The first few even harmonics are present because of asymmetry, due to the small dc component present in the current waveform [4]. From Figure 2, it can be seen that the current pulses are almost symmetrical with respect to $\omega t = \pi/2$ (for the positive half cycle) and $\omega t = 3\pi/2$ (for the negative half cycle). Therefore, the harmonics tend to add rather than mutually cancel [5].

From harmonic tests carried out on locally-available CFLs, using a Fluke 39 Power Meter, the following results were obtained:

Table 1: Summary of results obtained for CFLs tested

Range of THD_I	115.7% - 130.9%
Range of Power Factor	0.55 - 0.61
Range of Crest Factor	3.44 - 4.14

These high values of harmonic distortion suggest that the ballasts have very little filtering to attenuate the harmonic content. The low power factor of the lamps is due to the highly distorted shape of the current waveform. For non-sinusoidal signals, the power factor can be calculated by: [6]

$$PF = \frac{1}{\sqrt{1 + THD_I^2}} \cdot DPF \quad \dots(1)$$

where:

THD_I – total harmonic distortion in the current waveform

DPF – displacement power factor i.e. the angle of displacement of the fundamental component of the current to the voltage.

2.2 Effects of CFL harmonics on the distribution network and equipment

Effects on voltage distortion

Whilst the input power of a single CFL is quite insignificant (9W to 33W), about 3000 residential customers using a few CFLs per household, could easily create a total non-linear load of 300kW to 500kW. In a study carried out by Pileggi *et al* [5], it was found that for a 15kV class feeder with a maximum 10MVA load (which may realistically supply 3000 to 4000 residential customers), even two or three CFLs per household may be sufficient to cause voltage distortion in excess of 5%. However, in a similar study carried out by Dwyer *et al* [7], it was found that if CFLs with lower current distortion levels (30% or less) are used, voltage distortion problems will not occur on the feeder, even at the high penetration level of 150W of CFL per household.

Effects on transformers

The main effect of power system harmonics on transformers is the generation of excess heat due to increased losses that occur within different components of the transformer. The primary loss components are the winding I^2R losses, the winding eddy-current losses and stray losses from electromagnetic flux in areas such as the windings, core, clamp assemblies, and tanks.

Tests were carried out measuring the power flow into and out of a unity-turns ratio, isolation transformer, connected to an incandescent load and a similar CFL load. The results of these tests are found in Table 2.

From these results, it can be seen that more than 50% extra apparent power has to be supplied to the transformer to sustain the CFL load, as compared to the incandescent load, due to the low power factor of the CFLs. When this is considered in terms of a real feeder supplying a large CFL load, the transformer rating would have to be increased to prevent damage to the insulation of the transformer. The extra current supplied also implies derating the current-carrying capacity of the conductors, to handle the load. However, these factors are offset by the fact that for a given level of illumination, the CFL load will only be about a fifth of the incandescent load.

Table 2: Transformer test results for 100W incandescent and 100W CFL loads

PRIMARY SIDE OF TRANSFORMER			SECONDARY SIDE OF TRANSFORMER		
	100W incand	100W CFL		100W incand	100W CFL
P [W]	118	122	P [W]	93	91
S [VA]	126	191	S [VA]	94	178
PF	0.93	0.64	PF	0.99	0.51
I_{RMS} [A]	0.58	0.84	I_{RMS} [A]	0.43	0.73

Effects on circuit-breakers

In the presence of severe harmonics, circuit-breaker failure has been attributed to the inability of the blow-out coils to force an arc into the arc chute [8]. Furthermore, the extra heat dissipation that occurs in the solenoid of a thermal magnetic circuit-breaker, raises the temperature of the circuit-breaker, which lowers the trip point.

In tests carried out on a 1A thermally-actuated, miniature circuit-breaker, the tripping time was not significantly affected by the presence of a 60% third harmonic component. This is shown in Table 3 below.

Table 3: Trip times observed when testing MCB with fundamental and third harmonic

RMS CURRENT [A]	TRIP TIME WITH ONLY FUND. [s]	TRIP TIME WITH FUND. + 60% 3 RD HARMONIC [s]
2.3	32.5	30.2
2.6	24.2	24.8
2.8	21.3	22.9
3.1	18.7	19.0
3.3	16.0	17.1
3.8	12.7	11.9
4.0	9.9	9.5
4.1	0.01 (inst. Trip)	0.01 (inst. trip)

Effects on revenue meters

The induction disc watthour meter has been the most commonly used registration meter in residential and commercial applications. However, these meters were designed specifically to monitor only fundamental current. In the presence of harmonics, the torque produced on the disc by the electromagnet decreases, leading to discrepancies in the metering [9].

In tests carried out on electronic pre-payment meters installed by Eskom, it was found that these meters only register real power drawn. However, due to the low power factor of CFLs, they draw a significant amount of reactive power, which will go unregistered. This could lead to a loss of revenue from Eskom's perspective, unless a tariff is devised to compensate for the reactive power drawn.

3. ECONOMIC ANALYSIS OF CFLs

It is important to carry out an economic analysis of CFLs because it assesses the potential savings that could be incurred by using these lamps. In order to do this the following assumptions are made:

- The cost of a CFL is R70.00.
- The cost of an incandescent lamp is R3.00.
- The electricity tariff is 30c/kWh.
- The rated lifetime of a CFL is 10 000 hours, whereas that of an incandescent lamp is 1000 hours.

A comparison between an 'ordinary' incandescent lamp and a CFL is carried out, as shown in Table 4, using undiscounted values.

Table 4: Savings of Electricity and Money over the lifetime of a CFL

LAMP	CFL	INCANDESCENT
Wattage	20W	100W
Quantity	1	1
Hours Burnt	10 000	10 000
Lamp Life (Hours)	10 000	1 000
Energy Used (kWh)	200	1 000
Electricity Costs	R60	R300
Lamp Price	R70	R3
Total Cost	R130	R303

From the above table, it can be seen that a total of R173 would be saved if a 100W incandescent lamp were replaced by a 20W CFL. Also 800kWh of electricity would be saved over the lifetime of the CFL.

Using a similar analysis, the power consumption of a 20W CFL is compared to that of a 100W incandescent lamp, as used for a different number of hours per day for a year. The results of this analysis are shown in Figure 4.

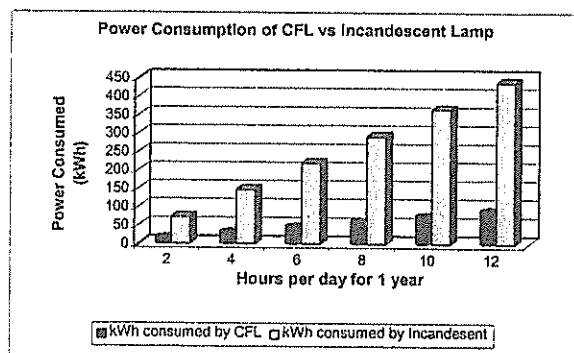


Figure 4: Relative power consumed by a CFL and an incandescent lamp of same lumen output

It can be seen from Figure 4 that the power consumed by a CFL is minimal as compared to that of an incandescent lamp. For example, if the CFL is used for 8 hours a day for one year, it would consume about 60 kWh as opposed to 290kWh consumed by an incandescent lamp.

Apart from the energy savings incurred, retrofitting an incandescent lamp with a CFL provides the further benefits of avoided costs in replacing incandescent lamps. This is due to the CFL having a longer rated life span than an incandescent; nearly 10 times longer. Taking into account the energy tariff and the avoided costs of replacing the incandescent lamps, the total costs of using a CFL in place of an incandescent lamp can be calculated. This is shown in Figure 5.

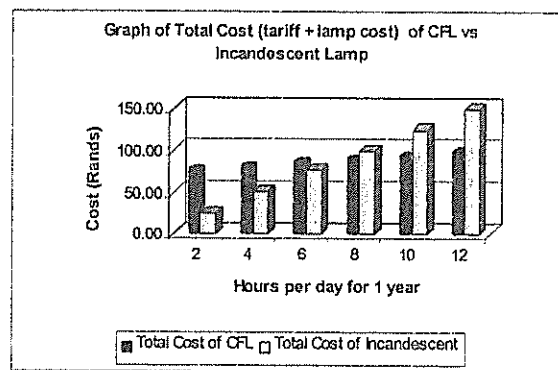


Figure 5: Relative costs of using a CFL and an incandescent lamp of same lumen output

The payback period of a CFL is calculated to occur at around 2850 hours. From the above graph, it can be seen that this corresponds to using the lamp for about 8 hours a day for 1 year. Alternatively, the CFL pays for itself if used for 4 hours a day for 2 years (neglecting the effects of discounting).

Apart from actual monetary savings, a further benefit can be accrued from using CFLs, in terms of avoided costs of environmental damage. The increased incidences of global warming and environmental pollution, as well as a decrease in the availability of our natural resources, is placing pressure on using environmentally-friendly technology. Since South Africa's demand for electricity is increasing rapidly due to the massive increase in housing development, Eskom can play a major role in reducing the impacts of power generation on the environment.

In a study carried out by Eskom [10], provisional estimates of the pollution that would be saved in South Africa per CFL, over its lifetime, were calculated. This is shown in Table 5.

Table 5: Estimated Pollution and Resources Saved per CFL during its Life Span

	CFL Power Rating			
	9W	15W	20W	23W
Water	351l	586l	781l	898l
CO ₂	260kg	433kg	577kg	663kg
SO _x	2.11kg	3.52kg	4.70kg	5.40kg
NO _x	1.06kg	1.76kg	2.35kg	2.70kg
Particulate	0.18kg	0.31kg	0.41kg	0.47kg
Ash	36.3kg	60.4kg	80.6kg	92.6kg

The table above shows that a considerable reduction in pollutants can be attained through the use of CFLs. These figures become truly significant considering that Eskom intends on installing 18 million CFLs throughout South Africa. [10]

4. CONCLUSIONS

Electronically-ballasted CFLs will play a major role in Eskom's RDSM programme because they consume only a fifth of the real power consumed by incandescents, and because of their high efficiencies. However, due to the distorted current that they draw, large CFL loads may jeopardise the quality of power supplied on the national grid, and may require upgrading of distribution equipment to handle the harmonics produced.

Although the initial costs of CFLs are high, it can be seen that these costs are recovered in the long run. The payback period of a 20W CFL is 2850 hours, which corresponds to 8 hours a day for 1 year. Therefore, purchasing such a CFL would save the consumer the cost of purchasing three 100W incandescent lamps and allow for a further 7 000 hours of use.

The energy efficiency of a CFL holds benefits for both the consumer and Eskom. The consumer is able to save electricity, and money in the form of decreased bills. Eskom benefits by avoiding the cost of installing extra generating capacity due to the reduction of electricity usage in peak demand periods. Eskom also benefits by being able to delay or prevent the upgrading of distribution infrastructure. A further advantage of using CFLs is the lowering of the levels of pollutants emitted, as a result of producing less electricity.

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PRELIMINARY FINDINGS ON DOMESTIC LOAD AND LOSS LOAD FACTORS

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Abstract. An approach has been derived to estimate the domestic network electrical losses over the system lifetime, using the load and loss load factors. This paper presents the preliminary load and loss load factor results, which have been derived from the available load data for 11 populations sampled from within South Africa. This paper introduces the most important concepts, outlines the problems involved, and presents some of the derived results. The derived factors are quoted on either a monthly or an annual basis. The approach used is based on a statistical sample from the available load data, considering small subsets within the population, and the population as an entity. The methods employ average values over long intervals, with a long-term view to the system electrical losses on either a monthly or annual basis, as appropriate.

Keywords: load factor, loss load factor, load data

1. INTRODUCTION

The accurate estimation of the financial implication of electrical losses in domestic distribution networks enables the designer and local supply authority to optimise the networks around the lowest life cycle cost, rather than the lowest initial cost.

This paper presents an approach used and preliminary results in the derivation of the load and loss load factors, which forms a portion of the analysis in the derivation of the system lifetime costs.

2. LOAD DATA CHARACTERISTICS

The available electrical load data consist of tables of values. The left-most column records date and time, with the stamped value recording the centre-most instant in a 5- or 6-minute period, depending on the local supply authority sampling standards. Attached to each of these date / time values are a series of logger channels with RMS current for a single residential consumer recorded in each channel.

Load data are bound between zero and the consumer circuit-breaker size, which can be as great as 80 A in some cases.

The approach to the analysis for load and loss load factors is based on the long-term thermal conditions of the distribution network. The time constants for the cables, which is the portion of the network under consideration, are typically considered to be approximately 30 minutes. Therefore the load data sample averaging intervals of either 5 or 6 minutes are acceptable.

3. LOAD DATA ANALYSIS

There are errors within the load data samples, relating to hardware and software failure. These are removed generally according to the methods described by Sellick and Gaunt [1]. These methods have been modified as the authors' understanding of the specific nature of electrical load data and their associated errors grew, and on a trial and error basis, comparing known real-world parameters to the derived equivalent values.

It should be noted that this error identification and removal is confined to entire channels, and does not focus on the errors that arise within a valid channel. This is an area requiring further investigation.

The "first-order" analysis that is required is the synchronisation of the loggers, to allow comparison across loggers, across time. This must also be sensitive to out-of-bound readings (typically negative, or close to 100A).

The methods used to sample sub-sets from within the sampled population are discussed under Section 0.

4. LOAD AND LOSS LOAD FACTOR

The load factor (LF) is defined as the ratio of the average load to the maximum load. The loss load factor (LLF) is defined as the ratio of the average losses to the maximum losses.

The relationship between the load factor and the loss load factor is governed by the relationship shown in Equation (1), and depicted graphically in Figure 1 [2].

$$0 \leq LF^2 \leq LLF \leq LF \leq 1 \quad (1)$$

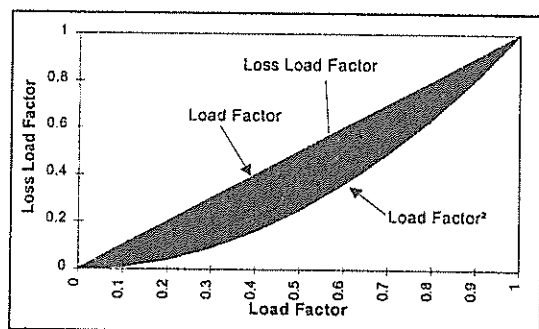


Figure 1 : Relationship between LF and LLF

The theoretical descriptions of population load and loss load factors, which are governed by the central limit theorem, indicate that these factors tend towards a theoretical limit as the number of consumers in the sample is increased. Similarly, general statistical theory states that the load factor for a number of consumers, $N + 1$ will be greater than or equal to the load factor for a number of consumers, N . This also applies to the loss load factor. These relationships may be expressed as:

$$LF(N + 1) \geq LF(N) \quad (2)$$

$$LLF(N + 1) \geq LLF(N) \quad (3)$$

The above equations are compared with the derived results below in Section 0.

5. ANNUAL vs. MONTHLY FACTORS

For the complete analysis, the load data has been considered on a "per month" basis, and on a "per year" basis. In each of these approaches, the load and loss load factors are considered for calendar months or years only.

Due to the total possible number of combinations of consumers from a population of, say, 70 consumers, a sampled sub-set has been chosen. This was set at 10 consumers. Although this may appear too few, only a small number of samples are required as the graphs are known to increase with n (ie. load and loss load factors always increase with number of consumers). The end points are fixed (ie. for 1 and for 70 consumers), and 10 possible samples provides an adequate representation of the link between these two extremities.

In each case, the values are stated as being "annual" or "monthly", although this is not strictly true. A 1997 annual graph refers to all values for the population within 1997, and a December 1996 monthly graph refers to all values for the population within December 1996. In practice, it is possible that there are only a few values recorded within a particular period quoted. This is particularly the case at the starting and ending points of the periods.

6. LF AND LLF VS. N

Graphs from one of the populations (Helderberg, 1997) are shown in Figure 2 to Figure 5.

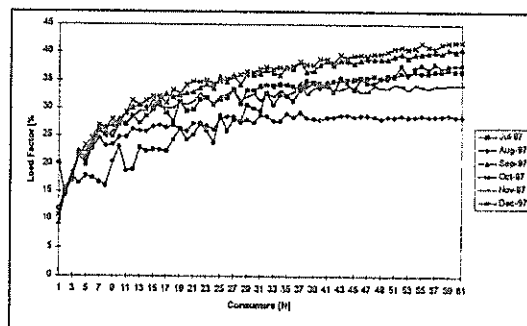


Figure 2 : Monthly Load Factor [%]

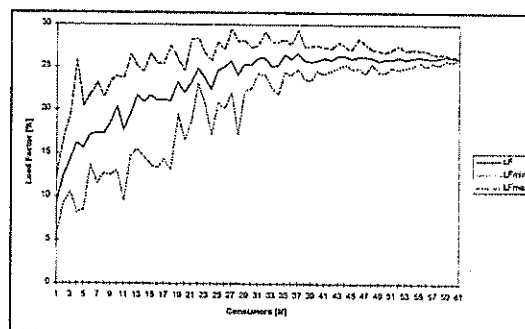


Figure 3 : Annual Load Factor [%]

The following important results are evident from inspection of these graphs:

1. Upon inspection of
2. Figure 2 to Figure 5, it can be seen that the practically derived results confirm the above theory contained in Equations (1) to (3).
3. The shape of the trend lines for the various individual months for both load factor and loss load factor are similar; there is a scaling factor which will depend on socio-economic status, climate and location, amongst others.

1. It is of importance that the graphs in each case tend towards a value, which is represented by the factor for a theoretically infinite number of consumers. This is not practically possible, but the population sample size (typically ± 70) has been chosen so that it is a sufficiently large sample size for the population under consideration.
2. All the populations that are reasonably error-free exhibit the trends exhibited by these populations in terms of generalised shape.

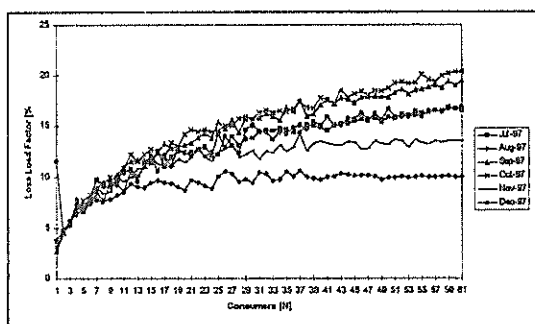


Figure 4 : Monthly Loss Load Factor [%]

In the monthly curves, only the average of the 10 samples taken has been plotted. In the annual curves, the three lines shown represent the maximum recorded value for the 10 samples, the average value, and the minimum recorded value for the sample of n consumers, where n ranges between 1 and the maximum population sample size available.

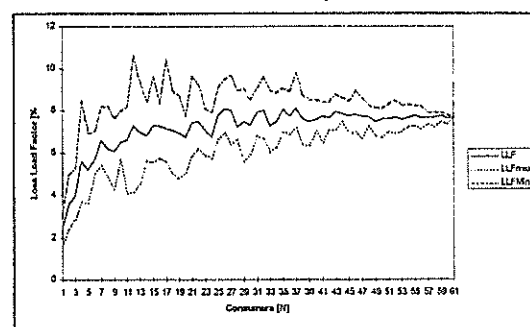


Figure 5 : Annual Loss Load Factor [%]

7. OVERALL POPULATION LF AND LLF

A summary of the maximum load factor and loss load factor available from the load data (due to the limit on the sample size) is shown in Table 1. The values are grouped within any one calendar year, and are quoted as a percentage. Some load data samples being at the end of one year and finish at the start of another year; readings within such a sample are grouped according to calendar year.

Note that in each case, the condition that the loss load factor lies between the load factor and the load factor squared is not violated. This is demonstrated in Figure 6, which plots the loss load factor for each year of each population, relative to the load factor and load factor squared.

The population / year no referred to in Figure 6 are the same as those in Table 1, and the values from

#	Population	Load Factor (LF)	Loss Load Factor (LLF)
1	Claremont 1996	1996 : 37.8	1996 : 16.3
2	Claremont 1997	1996 : 43.7	1996 : 23.5
3		1997 : 27.3	1997 : 12.2
4	Cloeteville 1994 / 95	1994 : 26.5	1994 : 10.2
5		1995 : 27.0	1995 : 8.6
6	Helderberg 1997	1997 : 26.1	1997 : 7.6
7	Kwazakhele 1995 / 96	1995 : 24.6	1995 : 9.2
8		1996 : 13.7	1996 : 7.3
9	Manyatseng 1996	1996 : 20.0	1996 : 10.3
10	Manyatseng 1997	1996 : 15.1	1996 : 3.0
11		1997 : 8.1	1997 : 5.5
12	Sweetwaters 1996	1996 : 10.5	1996 : 8.3
13	Sweetwaters 1997	1996 : 4.6	1996 : 4.6
14		1997 : 14.8	1997 : 8.7
15		1998 : 31.8	1998 : 10.5
16	Umlazi 1995 / 96	1995 : 38.2	1995 : 18.2
17		1996 : 34.8	1996 : 16.1
18	Walmer Estate 1997	1997 : 23.5	1997 : 9.3

Table 1 : Summary of Derived Load and Loss Load Factors [%]

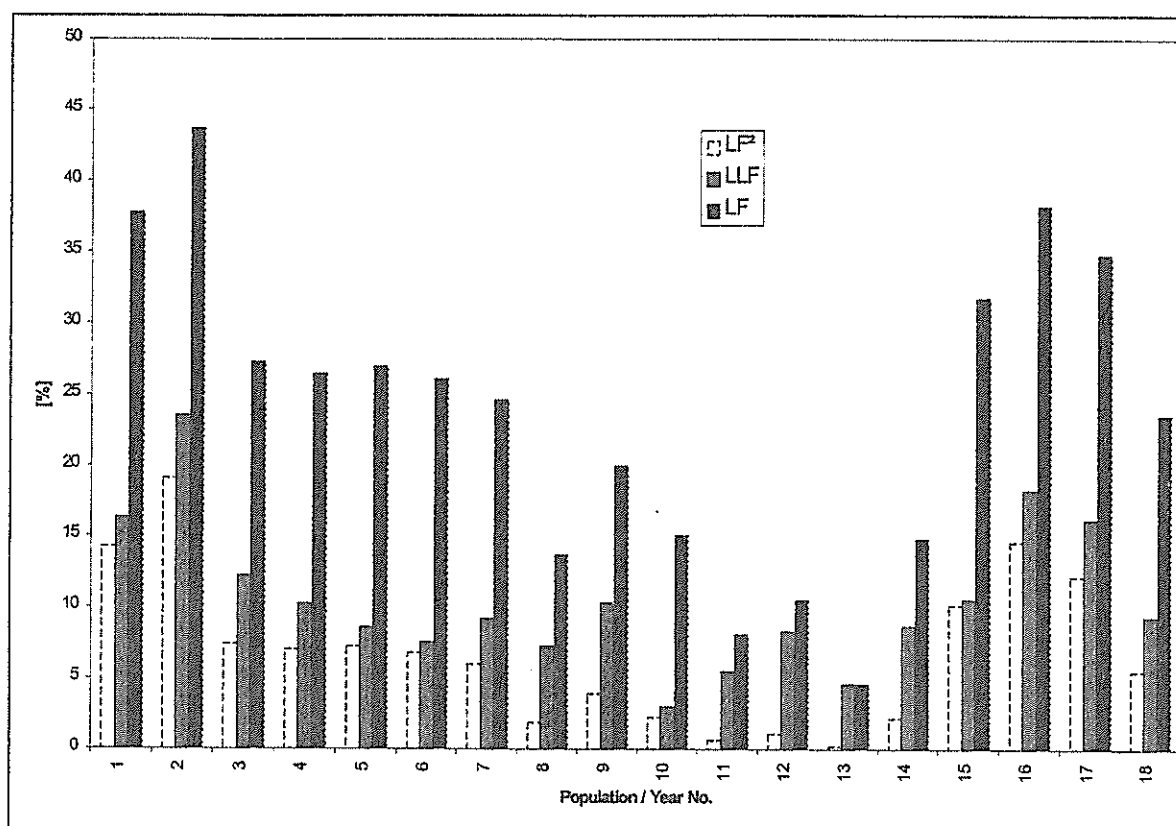


Figure 6 : Loss Load Factor vs. Load Factor

Table 1 have been used to plot the graph.

From Figure 6, it is of interest to note that for lower values of load factor, the loss load factor is relatively closer to the load factor than for higher load factor values, when the loss load factor is relatively closer to the load factor squared.

8. CONCLUSIONS

The factors derived are useful statistics for accurately describing specific electrical characteristics for the populations under consideration.

While the load and loss load factors are subject to external factors which govern the use of electricity, such as the socio-economic status, climate, etc, the general shapes of the derived curves for load factor and loss load factor confirm existing theory. These factors are functions of the consumer size, and must be treated as such when isolated feeders are considered.

The variability of the load and loss load factors requires further research, as these parameters change between different populations and in different

periods. The key parameters which govern these factors need to be identified.

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10. AUTHORS

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COST-EFFECTIVENESS OF THREE-, BI- AND SINGLE PHASE FEEDERS SUPPLYING DOMESTIC CUSTOMERS

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Abstract This study examines the costs of three-, bi- and single phase reticulation (distribution) on low voltage feeders supplying domestic customers. In the past, South African reticulation has been predominately three phase. Using new technologies available, the cost per erf can be greatly reduced. Several models are studied, from bi- phase reticulation is shown to be an attractive alternative.

Keywords: three-, bi-, single- phase reticulation

1. INTRODUCTION

There has never been as great a demand for electricity in South Africa as we are currently facing. With the implementation of the Reconstruction and Development Program (RDP) there have been millions of Rands invested into supplying new households with electricity. As a result, there is a need to assess the method by which electricity is delivered to the customer.

Reticulation (distribution) developments last occurred from the 1940's till the late 1950's. Since then, reticulation has been very much standardised. With the introduction of new technologies, the potential to reduce the cost of electrification is made possible. This study studies 230V single phase, 230/460V bi-phase and the conventional 400V three phase feeders.

The objective of this study is to assess the three above mentioned forms of reticulation in terms of cost, using cables and transformers of various sizes.

2. SCOPE OF THE STUDY

In order to compare the three proposed systems, models are developed based upon three customer densities and loads. The models to be investigated are the high density communities with low demands, low density communities with low demands, and low density communities with high demands. The models are as follows for the two densities:

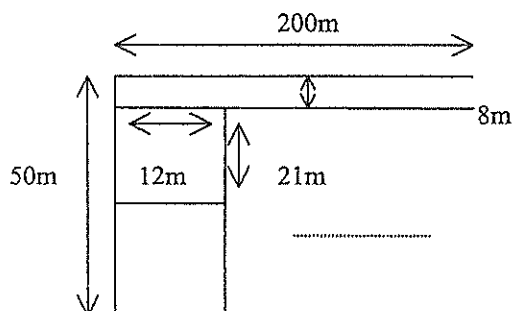


Figure 1: Layout of the 32 dwellings/ha

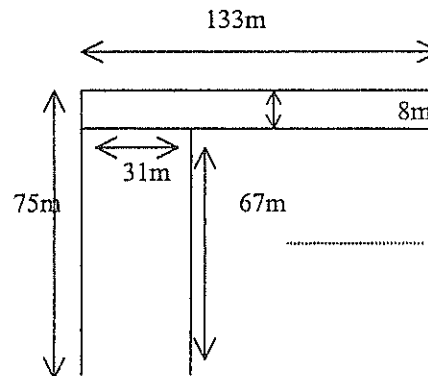


Figure 2: Layout of 4 dwellings/ha

This study assesses the cost implications of alternative technologies for supplying domestic customers using theoretical and practical Aerial Bundle Conductors (abc). Theoretical conductors are conductors are not constrained by manufacturing standards. On the other hand, practical conductors are ones which are manufactured as a standard.

The costing includes all the elements of the low voltage (LV) reticulation network, including the medium voltage/low voltage (MV/LV) transformer up to, but excluding the customer service connections. Only 10kVA, 25kVA, 50kVA, 100kVA and 200kVA pole mounted transformers are considered. The basis of the costs estimated are consistent for all the cases investigated.

3. METHODOLOGY

The models allow for several feeder typologies to be investigated as well as their effects on load and voltage conditions. Beta [1] parameters are calculated for the three models. Then, using the Herman Beta algorithm in the form of a spreadsheet [2], voltage conditions on the feeders are calculated. The theoretical cable size is selected in order to give a six percent voltage drop [3]. In order to calculate the practical cable size, the theoretical cable size is simply rounded up to the closest manufactured standard. With this information, a schedule of costs can be proposed for each model.

4. RESULTS

The results are presented for the three density-load models. In all the cases the practical results follow the same trends as the theoretical results presented.

4.1 HIGH DENSITY, LOW DEMAND COMMUNITIES

Figure 3 displays the results of the theoretical cable calculations. The results indicate that dual phase is the most economical form of reticulation from 10kVA to 160kVA transformers. From 160kVA to 200kVA it is more economical to implement a three-phase network.

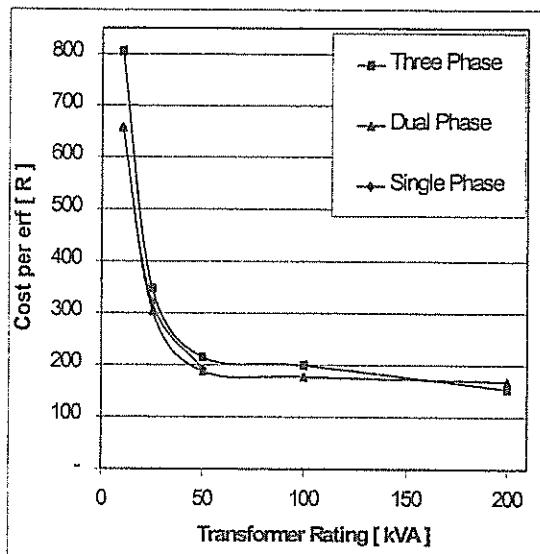


Figure 3: 32dwellings/ha & ADMD of 1kVA. Theoretical Results

4.2 LOW DENSITY, LOW DEMAND COMMUNITIES

Figure 4 indicates that for this specific load and density any transformer larger than 50kVA is not viable. In this model, bi-phase reticulation is the most economical.

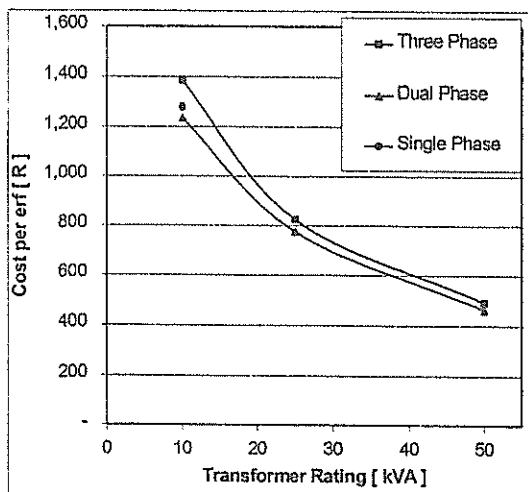


Figure 4: 4 dwellings/ha & ADMD of 1kVA. Theoretical Results

4.3 LOW DENSITY, HIGH DEMAND COMMUNITIES

In Figure 5 a single-phase system is the most cost-effective means of supplying electricity between 10kVA and 20kVA. From 20kVA to 100kVA bi-phase reticulation is the most economical with an average saving of 10% compared to a three phase system.

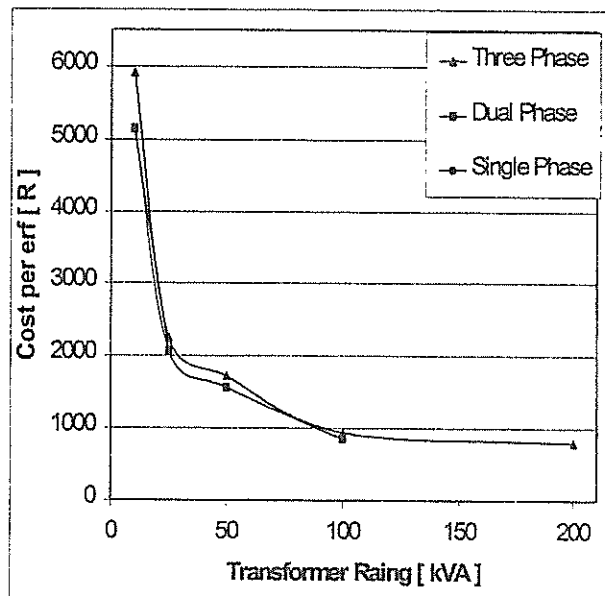


Figure 5: 4 dwellings/ha & ADMD of 4.6kVA. Theoretical Results

5. DISCUSSION

5.1 HIGH DENSITY, LOW DEMAND COMMUNITIES

The lowest cost/connection for this model is achieved with three-phase reticulation at 200kVA. It must be recognised, though, that the model is based on rectangular erven as indicated in section 2 and makes no provision for open spaces, schools, etc, that occur in practise.

Bi-phase LV feeders from transformers rated 10 to 160kVA result in costs significantly less than three-phase reticulation over this range, and a cost/connection almost as low as the three-phase cost/connection for a 200kVA transformer. Bi-phase reticulation is on average 19% cheaper than the three-phase for 10 to 160kVA transformers. The network layout in practical communities is likely to be more efficient using smaller transformers and bi-phase feeders instead of three-phase systems.

5.2 LOW DENSITY, LOW DEMAND COMMUNITIES

Bi-phase reticulation for LV feeders incur the lowest cost of the three technologies over the entire range of transformers up to 50kVA. Above 50kVA, the size of the abc feeders becomes impractical in order to meet the specified voltage drop. This model needs to

be tested further, using larger transformer ratings with multiple LV feeders of practical cross-sectional area, to determine the break point at which the increasing volume and the cost of the LV conductor exceeds the savings in the transformer and MV network.

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5.2 LOW DENSITY, HIGH DEMAND COMMUNITIES

Bi-phase reticulation is the least expensive technology with transformers up to 100kVA rating, at which the size of the abc becomes impractical. The high load carrying capacity of three-phase reticulation makes it possible to use transformers with a rating of up to 200kVA for the LV feeder model used, with a similar cost/connection as dual phase with 100kVA transformers .

6. CONCLUSIONS

Single-phase LV distribution appears to be viable only for small dense communities. It is however unlikely that such high densities will occur in small isolated areas and therefore larger transformers offer lower costs/connection as communities grow.

Bi-phase reticulation appears to offer significant savings over three phase reticulation for a wide range of customer densities and load sizes. It is recognised that the structure of the model may introduce distortions of the actual costs, but the relative costs consistently indicate that dual phase feeders incur the least cost.

7. RECOMMENDATIONS

More detailed comparison of bi- and three-phase distribution using real layouts appears justified. The range of comparisons should include:

- ☐ Customer densities from 2 to 32 dwellings/ha;
- ☐ Transformers from 25 to 200kVA; and
- ☐ Multiple LV feeders where constrained by practical conductor size.

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Energy Management: Using the Durban and PMB campuses of the University of Natal as case studies

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Abstract: This paper deals with the investigation and the results obtained from looking at the way the campus utilize its mechanical resources. There is no record of energy usage and energy accountability. The sources that contribute to the load profile and the magnitude of contributions are unknown. This project sets out to solve the abovementioned problems.

Keywords: Energy Management, Energy Audit

1. INTRODUCTION

At the beginning of this year a project was undertaken to look into the energy usage on the campuses of the Durban and Pietermaritzburg campuses of the University of Natal.

This paper highlights the results of this investigation and proposes strategies that could affect cost saving. These strategies that could also be applied to any industrial or commercial venture that uses a substantial amount of electrical energy.

2. WHAT IS ENERGY MANAGEMENT

Energy management embodies engineering, design, applications, utilization, and to some extent the operation and maintenance of electric power systems to provide for the optimal use of electrical energy. Optimal in this case refers to the design or modification of a system to use the minimum overall energy where the potential or real energy savings are justified on an economic or cost benefit basis. Optimization involves factors such as comfort, healthy working conditions, practical aspects of productivity, aesthetic acceptability of space and public relations. The idea is to reduce levels of energy consumption without a consequent reduction in the quality of education and service provided. The audit was conducted using the following Fig. 1 as a guide.

Block 1 in Fig. 1 has been explained. To implement block 2, an energy database had to be established. There were no previous records, except for the energy bill, as to energy usage. The utility, which supplies the University, keeps a copy of the load profiles of all its customers. These profiles are the profiles of the usage as seen by the main incomer. Using these profiles, and determining the tariff structure on which the total bill was based on, it was confirmed that the Time-Of-Use (TOU) is the correct available tariff structure.

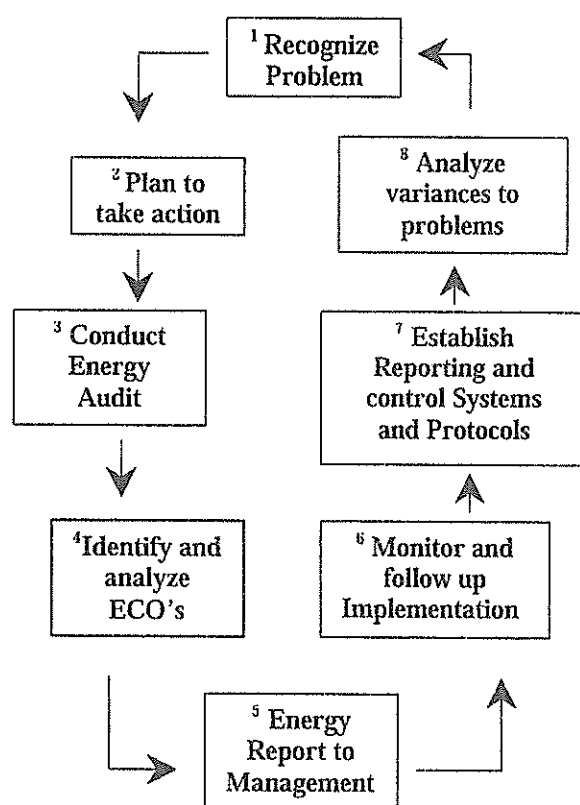


Fig. 1 Flowchart to conduct energy audit

The TOU tariff uses three billing periods during which the University is charged for the energy that it uses: peak, standard and off-peak. In addition to these energy charges there are also demand charges (which at present account for one third of the total bill). There are two types of demand charges: maximum demand and overall demand. The maximum demand charge is calculated

by looking at the peak demand in the billing month only during the peak and standard periods. The overall demand charge is calculated by looking at the peak demand during all periods.

By looking at the University's load profile from the main incomer it was immediately evident that the major portion of the energy was being consumed during the peak and standard periods. The peak overall demand was also occurring during these peak and standard periods. Therefore the University was paying two demand charges during the peak and standard period, one for maximum demand and the other for overall demand. A typical load profile is shown below :

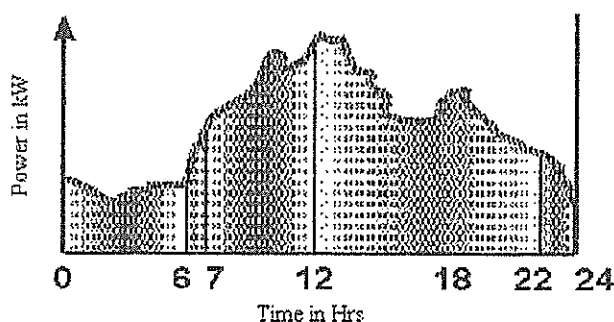


Fig. 2 : Typical Weekday profile from Main Incomer

For the month of September 1998 the energy bill was R 334 771.66. By modifying the shape of the above profile, Fig. 3 was obtained.

By consuming the same units of energy but with a load factor of 1, the bill would be R 277 964.42, resulting in a saving of R 56 807.24.

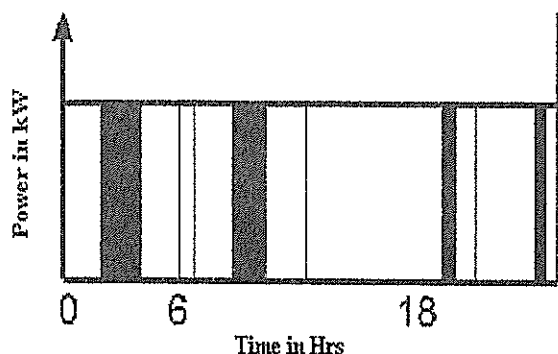


Fig. 3 : Flat Profile for Weekday in September 1998

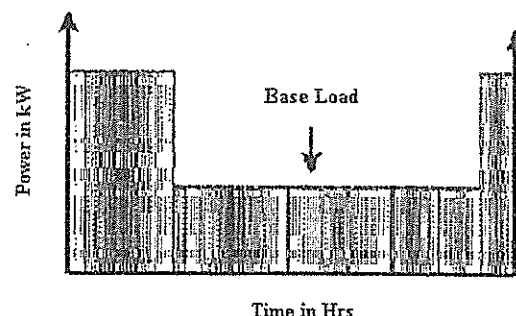


Fig. 4 : Optimum Profile for Weekday in September 1998

By further modifying the profile as shown in Fig. 4, the energy bill, based on the same units of energy, would be R 250 069.03 resulting in a saving of R 84 702.03. Fig. 4 was determined by only running the base load during the peak and standard times and running the rest of the load during the off peak period. This was found to be the optimum profile to maximize savings but is a theoretical curve. It may not be practically realizable due to system constraints but this is the curve to aim for. To be able to obtain this curve, all controllable loads need to be identified.

Metering is essential to any energy audit. Due to financial constraints meters were not installed on all eighteen transformers on the main reticulation system. The potential major controllable users of energy on campus were identified. Meters were borrowed from Eskom Elektrowise and installed on the transformers supplying the central air-conditioning system and the residences. Funds have finally been procured for more meters, which will be installed on the larger energy users on the reticulation system. Ultimately the aim is to establish and improve energy use and accountability for each department. A proposal would be to charge out energy to individual departments based on the TOU tariff. Metering would also help to determine the efficiencies of systems. Also power factor correction points will be identified.

The figures shown below are the profiles that were obtained from the meters that were installed on the transformers supplying the central air-conditioning system and the residences.

Albert Luthuli : Typical 10 Day Period (24th is Holiday)

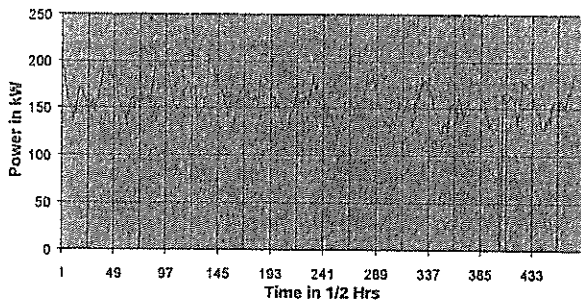


Fig. 5: Profile from Main Incomer in a Residence.

An examination of the above profile, shows that the major consumption to be during the evening, due to cooking. Savings can immediately be made by scheduling cooking periods a few hours earlier and warming foods when required for eating. Water heating is also a problematic area. It was found that the water tanks were being heated 24 hours a day. By raising the thermostat temperature 10 degrees higher, the tank could provide the same capacity and be charged during off peak periods only. This results in a saving of R41 121.36 per year on a single residence just by just rescheduling water heating. The University presently has a running kWh meter attached to the main incomer of each residence and the residences' budgets are debited on a pro rata basis based on these running meters. This is misleading because it does not take into account demand charges. A proposal is to put in individual card meters, which would work when a student card has been swiped and to debit students for usage and informing the students of suitable periods for cooking and warming foods. This could also be a possible solution to illegal squatting in residences.

The air-conditioning on campus for some of the larger buildings are done via a central chilled storage tank. The water in the tank is chilled to 4 ° C and pumped out to the thermal load requiring space conditioning. The return water should ideally have a return temperature of 14 ° C to ensure maximum storage capacity. This ensures a thermocline separates the chilled water from the heated water. The advantage of this system is that, chilled water can be stored to create some kind cooling capacity. The Thermal Storage System (TSS) is not a new concept but it is very effective. Unfortunately it is not as effective in the University TSS due to the problems mentioned below.

The chillers are deteriorating and are not running efficiently. The refrigerants are also not the most effective both from a cost and an energy perspective.

Due to the size of the chillers and the type of refrigerant used, the TSS has to go through two cycles to reach the required temperature difference of 10° C. The time taken to go through these two cycles and fully charge the tank exceeds the required eight-hour window of off-peak time. The entire concept of TSS is based on this window. The block diagram in Fig. 6 shows a single cycle.

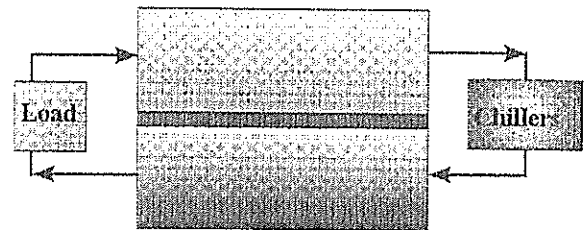


Fig. 6. Process of a single cycle to charge TSS tank.

Total control over the entire system is lacking. The presence of split units in various buildings also compounds the problem. There is no complete record of the number of split unit air-conditioners that are installed and operational. This is currently being undertaken. Their impact on the total energy usage will be soon known. There are also no operating schedules for the split unit air-conditioners. Some are known to be perpetually on, even at times when they are not required.

Chiller Consumption for 28/09 to 4/10

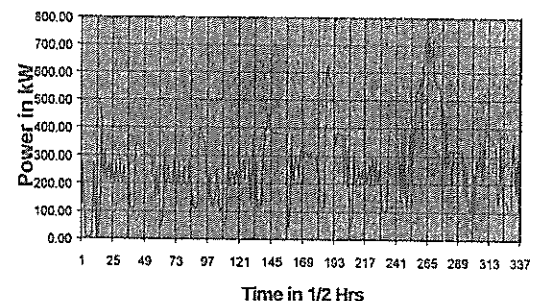


Fig. 7 : Profile from Transformer supplying Chillers

An examination of Fig. 7 shows that the thermal storage tank is being charged during peak and standard times on certain days. This practice not only incurs a higher energy charge but also increases the aforementioned demand charges. The power factor

of the TSS was measured to be 0.8. This is due to the amount of equipment connected to the entire system namely: pumps, compressors, induction motors etc. An improvement of this figure is currently being investigated.

The task that is currently being undertaken, is analyzing each building and obtaining a benchmark figure for energy consumption levels for a building of that size and then comparing the calculated figure with future measurement readings. Any deviation will have to be investigated and accounted for. By accomplishing this for all buildings on the University campus, the lower profile shown in Fig. 8 will be attained.

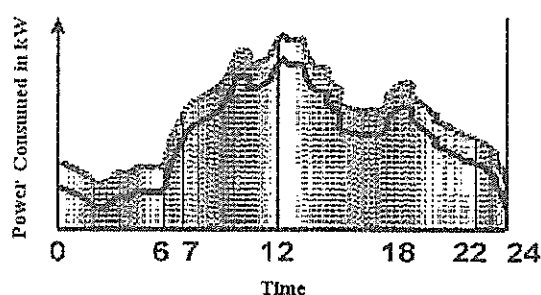


Fig. 8: Projected Profile from Main Incomer

The targeted figure is a saving of twenty percent using present levels of energy consumption. This is certainly realizable from results yielded by the initial investigation.

3. CONCLUSION

It has been found that there is no real awareness in terms of how present energy usage affects the total energy costs. Therefore there is inadequate or in some instances no housekeeping measures being taken. Equipment and process modification, in terms of the TSS, is currently being addressed. Better utilization of equipment is also being addressed. There needs to be an energy committee, with major players in terms of mechanical resources involved, to tackle the problem from a global perspective. As shown by Fig. 1, this is an on-going process that will have to continue after the completion of this project to sustain the savings made.

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Energy Management Study on the Potchefstroom University for Christian Higher Education Campus

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Abstract

Energy Management is an increasingly important issue for all Universities across the country because of budget cuts and increased financial strain. This article describes the problems and solutions encountered in a survey to establish the possibility of installing an active demand control program at the University.

Keywords: Energy Management, Load survey, Energy Strategy, Peak Demand

1. Introduction

The Potchefstroom University is located in the North West province of Southern Africa. What makes the University an ideal client for energy management is that the Universities electricity consumption is measured at one central point.

What complicates the study however is that energy management is a difficult issue since the comfort of all students has the highest priority. This implies active energy management without discomforting the user.

To achieve this, complicated control techniques, highly trained staff and advanced electronic equipment is required to achieve the task.

2. Energy Loads at the Potchefstroom University

The big loads on campus that were identified as possible energy saving equipment, and that might be causing peak demand problems are:

1. Hot Water Boilers;
2. Air-Conditioning Heating Elements and
3. Chillers and Fans.

The contribution of each of these loads are given in the following table.

Equipment	Rating
Boilers	1455 kW
Heaters	2050 kW
Chillers	582 kW

Table1: Contribution to the peak load

Hot Water Boilers

The Potchefstroom University took its first step towards active demandside management in 1990 by installing a system to measure and control the boilers used in the student hostels.

The system measures the temperature of each geyser and additionally logs the demand usage of the campus. If the kVA for that demand period comes close to the set maximum kVA of that month, the system starts switching of geysers, starting with the hottest one (this within set points).

The first fully operational year of this system was 1991, and since then the changes in electrical equipment and the growth in student numbers caused the demand to rise to previous levels. Figure 1 shows the total yearly kVA since 1987. A decline in the kVA usage can be seen when the system was installed, but a rise in the yearly load urged the University to look into new alternatives to save on their electricity bills.

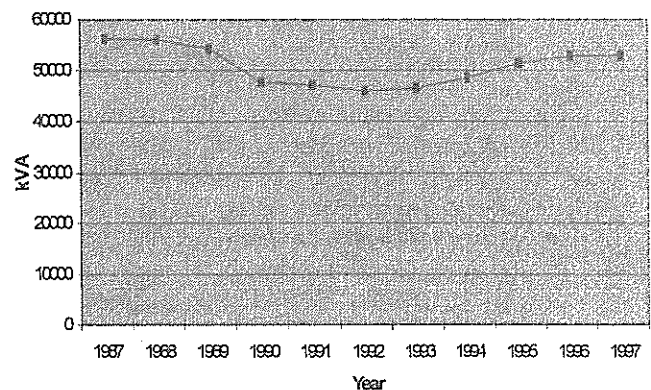


Figure1: Accumulated yearly KVA demand

Air-Conditioning Heating Elements

The heaters on the campus are some of the bigger individual loads. In the winter months they cause, together with the geysers, for a higher maximum kVA. The amount of kVA that can be saved by controlling these heaters depends on the frequency of use, and most important, comfort factors.

Chillers and Fans

The chillers are smaller loads but are used more frequently than the heaters. Like the heaters, controlling these chillers can be difficult due to the comfort factors that are involved. Currently the control of most of the chillers is very efficient. It only cools a lecture room when a start button is pressed and times out after 50 minutes. There is no pre-chilling of lecture rooms.

3. Campus Load Profile

Since the geyser control system was installed, total energy usage of the campus have been measured and stored on to discs. This data contains the daily values of kW, kVA, and the Power Factor with 320-second intervals, and was used for the calculations given in this paper. This data were used to determine:

1. Typical load curves of the university for different days and seasons;
2. The Loading Factor
3. When maximum demands occur, and;
4. Possible savings.

After integrating the values over 30-minute intervals (the integration period of the council meter), the values could be used for our calculations.

Total load curves

The total load curves of the university for a typical summer month are shown in figure 2.

The study indicated that the time and size of the peaks in the winter and summer are different. In the summer the main peak is between 10h30 and 12h30, which is later than in the winter where the peak stretches from 8h30 till 11h00. A reason for this is that, in the summer, more and more air-conditioning units are switched on as the day progresses, while in the winter the heaters are switched on early in the morning because of the cold. The higher peak in the winter can be explained by the larger energy consumption of the heaters and boilers, compared to the

chillers. This explains the effect heating and cooling has on the total profile.

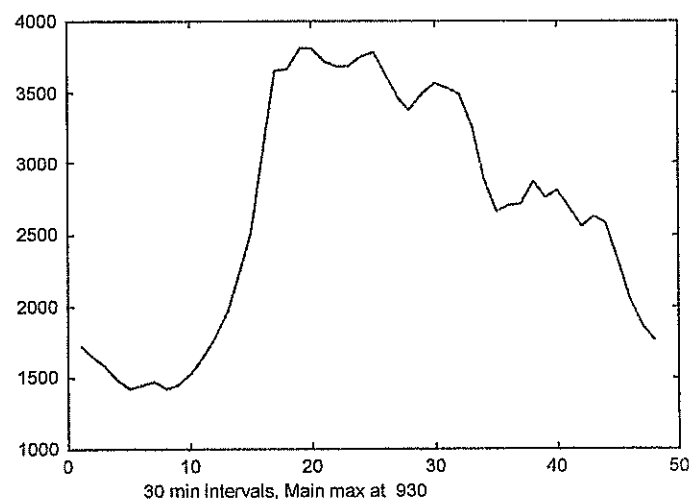


Figure 2: Campus load Profile (February 1997)

The second peak in the winter is between 19h00 and 22h00, which is caused when most students are in the hostels and they use warm water, heaters and lights.

Loading Factor

The loading factor calculates how evenly the electricity usage is spread over the day. If the loading factor is close to 1, it implies the load is spread evenly for the day, and no peaks occurred.

The loading factor of the university is shown in table 2 (only for Monday to Friday). The higher peak in the winter (June) results here in a lower loading factor.

Month	Average Loading Factor	Minimum	Maximum
February 1997	0.72	0.69	0.75
June 1997	0.70	0.63	0.77

Table 2: Loading factor of PU

Time of Maximum demands

Examining the total load curves, we could determine what time of the day was generally responsible for the peak of the month. To have a better idea what time of the day contributes the most to the maximum, we

determined in what integrated half-hour period the kVA value was the highest.

In the summer the intervals 11:00h - 11:30h and 12:30h - 13:00h caused the most occurrence of peaks, and in the winter the intervals between 9:00h and 10:30h.

Possible savings

To lower the maximum monthly demand of kVA by a certain percentage, you need to know how much to shed, how often and for how long.

If, for example, the desired saving is 3% for the month June, it was determined that the biggest peak above this set point was on the 3rd of June between 9:00h and 11:30h (note: the time indicates the end of the integration period in which the value was determined). This gives proof that the peak that occurs form a plato on the graph, stretches over a large part of the day. Estimated savings, % of available load and shed period, if the peak could be reduced with proposed techniques ,are given in table 3.

	Month	Savings (Rands)	% available load	Shed period(h)
5 % Saving	Feb	6583	19.2	3.5
	June	8713	26.9	2.5
	Nov	6164	24.2	5.0
	1997	85840	-	-
4 % Saving	Feb	5260	12.4	3.0
	June	6970	20.3	2.5
	Nov	4905	15.5	4.0
	1997	68540	-	-
3 % Saving	Feb	3937	7.8	2.0
	June	5228	13.7	2.5
	Nov	3679	6.1	1.5
	1997	51376	-	-
2 % Saving	Feb	2646	3.8	2.0
	June	3485	8	2.0
	Nov	2453	3.6	1.0
	1997	34336	-	-
1 % Saving	Feb	1323	1.2	1.0
	June	1743	2	1.5
	Nov	1226	1.8	1.0
	1997	17168	-	-

Table 3 : Possible Savings for different percentages of peak reduction

4. Electricity Usage of Groups and Buildings

To determine where the loads are which contribute the most to the peak electricity usage of the university, individual buildings or groups of buildings have to be measured. At the same time the total usage of the university needs to be measured, and the contribution to the peak of the building or group can then be calculated.

It is often not possible to measure the group of loads desired. This is caused by the physical layout of the electricity network and/or the absence of measurement points, suitable for a logging meter. With load surveys the absence of sub-metering is often a problem, as is the case. A suggestion was made to the university to install a permanent sub-metering system, which would consist of 11 three-phase meters, which will accurately desegregate the power usage for the campus. These meters can be used for more accurate sub-metering, and helping with the education of people towards energy savings.

The other big users are the boilers used in the hostels. To determine what effect the geysers have on the total load, one of the hostels was measured and normalised to represent the total boiler load on campus. In figure 3 the resulting load is shown after subtracting this calculated curve from the average total load curve of that week.

The result is a lowering of the first and second peak of that day, emphasising the part the boilers play in both load peaks. In this week the current control system of the geysers did not switch any geysers off because of the low demand in that period.

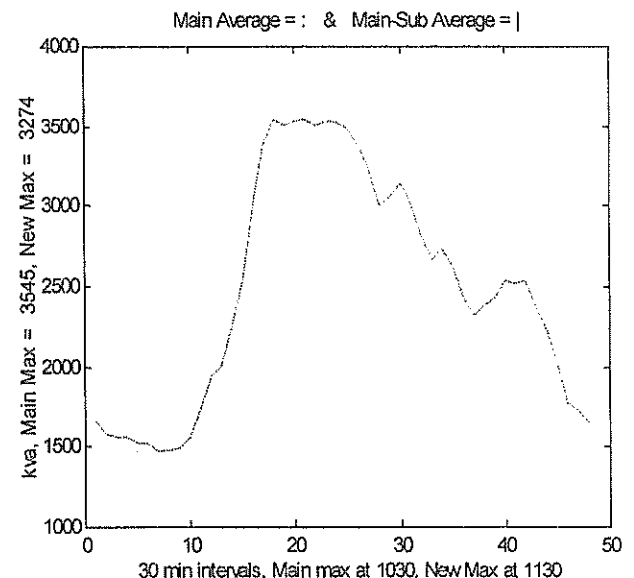


Figure 3: Resulting load curves

5. Energy Management Strategy

The following key issues were determined to be on the critical path for the successful implementation of an energy management program.

❖ Simulate Boiler Activity on Campus

The School of Electrical and Mechanical Engineering are currently busy with simulations and measurements to create a model to actively simulate different control scenarios on the boilers. As soon as such a model is complete it will be possible to simulate different control methods on the boilers, without influencing the comfort factor for the students involved.

In-Line heating techniques are also experimented with in order to try and change the effectiveness of the boiler.

❖ Sub-metering Project

A sub-metering project was launched to be part of the permanent operations on campus. The problem however is to justify the expenditure for a project that will result in no immediate financial gain to the university.

❖ Upgrading existing hardware

To generate additional savings to the current energy control system, advanced measurement and control techniques will need to be used. For this to be possible intelligent hardware control modules are needed to make control possible. A distributed control technology called LonWorks is to be used as the technology platform for further expansion on. Reasons for choosing this technology is:

- Lonworks supports an open architecture
- Lonworks implements a distributed control architecture and not centralised control like in the past.
- The University developed major expertise in the development and implementation of the LonWorks control architecture.
- LonWorks supports a huge variety of communications media, including internet connectivity.

❖ Evaluate Heating and Cooling on Campus

Lecturers from Mechanical Engineering were asked to determine if the air-conditioning of the lecture

halls are used as energy conservative as possible. They determined that the switching of equipment with timers as being done at present, are most effective and that further improvements does not validate the expenses in know-how and equipment that will be needed.

The in-line heaters used in the air-conditioning of buildings are switched on in stages, and is an effective way to regulate the amount of heat used. Experts need to determine what the influence on the comfort would be in places like the Library, if these loads were to be shed as part of the energy management program.

❖ Benchmarking of Buildings

Another issue that are being investigated is the benchmarking of buildings on campus by using the initiatives of personnel participating in the Green Buildings initiative.

Conclusion

The energy study highlights the different aspects when doing an energy survey. The survey has shown that energy savings at the University is possible, but that the process is also an education process to establish a culture of saving energy as much as it is an engineering challenge.

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7. About the author

Johann van Rensburg holds a B-Ing (Electric & Electronic) degree from the Potchefstroom University for Christian Higher Education. He is a part-time scholar at the University with the above project being part of his Masters Degree in Energy Management. He is a full time Employee of LonTech (Pty) LTD, which concentrates on LonWorks based control applications.

CONDITION MONITORING OF INDUCTION MOTORS USING FLUX COIL

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Abstract: This paper attempts to identify induction motor faults which can be detected by a leakage flux spectrum analysis. Tests were performed on squirrel cage induction motors demonstrating faults such as unbalanced supply voltage and stator winding faults. Methods of unbalanced voltage supply and stator winding faults detection are described in detail and example results presented.

Keywords: condition monitoring of induction motors, flux coil, leakage flux monitoring, induction motor faults

1 INTRODUCTION

Over the last years different techniques have been developed for the early detection of induction motor failures. Although the amount of information they provide regarding the machine condition is reliable, they have several limitations i.e. stator and rotor failures must be detected off-line and with the machine dismantled. This will mean complete shutdown of the plant for critical motors in the plant e.g. boiler feed pumps in power station. In order to overcome these shortcomings researchers in this field of condition monitoring of electrical machines made attempts to find non-invasive method based on the parameters which will allow early detection of failures in these machines.

In the ideal motor the axial leakage flux is zero. This is because under fault-free conditions, the rotor and stator currents should be perfectly balanced. However, in practice asymmetries exist in both circuits due to slight imperfections in winding geometry as well as non-uniformities of the magnetic circuits. The net effect of these deviations from the ideal motor is the production of small but measurable axial leakage flux. The premise underlying the use of axial leakage flux as a condition monitor is that faults represent large asymmetries in the machine windings. Hence the effect of a fault will be to enhance the magnitude of the axial leakage flux. From a relatively simple analysis, the harmonics of the axial leakage flux can be derived. In addition the harmonic components of interest, generated by a particular fault can be, in most cases, predicted.

The flux coil is a simple sensor made of a copper wire. The more turns the larger the signal. Normally 200 to

300 turns is sufficient. When wire is wrapped to form a coil the resulting flux coil is fairly flexible. Therefore, the coil can be shaped to fit in and around many odd shaped spaces. However, in order to collect repeatable and reliable flux data, consistency is a major concern. Even so a measurement can be taken at the same place of the motor, if the coil shape is significantly different from one measurement to the next the magnitudes of the flux harmonics can vary. This limits the reliability of the data. Therefore, a formed flux coil is most desired. Consistent placement of the flux coil on the axial outboard end of the motor is critical for obtaining reliable data. It is important that the measurements be taken at the same location, with the same spacing between the sensor and motor, and without swinging or twisting movement of the flux coil.

2 UNBALANCED VOLTAGE SUPPLY

In electrical installations where the supply system is relatively small then the 3-phase voltage supply could become unbalanced due to unbalanced single-phase loads. This causes overheating of the windings, thermal ageing occurs and can results to insulation failures, insulation life can be halved for every 10 °C rise over the rated temperature [1]. Numerous protection schemes are available for preventing motor failures due to unbalanced supply and are usually based on sensing over-current, voltage or current asymmetry or stator winding temperatures. However, they are not always successful and motors still fail due to the effects of unbalanced supplies [1,2,3].

The effect of unbalanced supply voltage in induction motors has been researched and researchers have found that this type of fault produce unique harmonic

frequencies in axial flux spectrum [1,2,3]. The predicted flux harmonic frequencies are discussed in the next section.

2.1. Predicted harmonic frequencies in the flux spectrum due to supply voltage imbalance

Unbalanced supply voltage results in lower value of current in a phase which leads to phase asymmetry in the stator winding. According to researchers in [3], such condition introduces components of flux density given by:

$$B = \frac{1}{2} \sum_{n \rightarrow \text{odd}} \hat{B}_n [\cos(\omega t - np\theta) + \cos(\omega t - np\theta)] \dots \dots \dots 1$$

which when transferred to the rotor frame of reference, becomes

$$B = \frac{1}{2} \sum_{n \rightarrow \text{odd}} \cos[(1 \pm n(1-s))\omega t \pm np\psi] \dots \dots \dots 2$$

The components of equation 1 induce currents at frequencies given in equation 2. These frequencies are:

$$f = (1 \pm n(1-s))f_s \dots \dots \dots 3$$

where n is an odd integer.

A change in amplitude of these frequencies indicates the asymmetry in the stator winding. Therefore these frequencies can be used for the detection of unbalanced supply voltages. The predicted harmonic frequencies using equation 3 for the test motor are given in Table 1. Researchers also report [1] that frequency components for $n=1$, i.e. $(2-s)f_s$ showed significant change in the flux and vibration spectrum. The change in vibration spectrum as a result of voltage imbalance is explained as follows. During normal operation of a 3-phase induction motor it is reasonable to assume that the magnitude of the fundamental air gap flux remains constant between no-load and full load operation. This is true from the basic equation of e.m.f. for a three phase motor i.e. $E = 4.44fN\phi k_w$, where N , f , k_w are constant for a given motor design. Change in applied voltage results in the change in the flux density and the magnetic forces which act on the stator core are proportional to flux density squared. Then any change in the flux distribution will be observed as the instantaneous change in the vibration spectrum [1].

The effects of unbalanced supply on axially directed leakage flux was measured using a 380 V, 4-pole

Table 1: Predicted flux harmonic frequencies due to unbalanced supply voltage

Frequency multiplier	Slip	Supply Frequency Hz	Harmonic Frequency Hz
s	0.024	50	1.20
2 - s	0.024	50	98.80
3s - 2	0.024	50	96.40
4 - 3s	0.024	50	196.40
5s - 4	0.024	50	194.00
6 - 5s	0.024	50	294.00
7s - 6	0.024	50	291.60

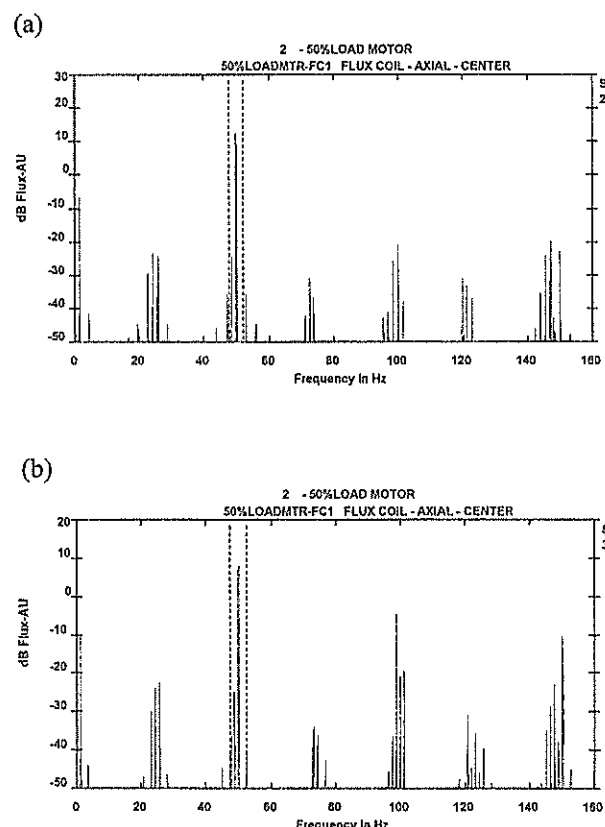


Fig. 1. Leakage flux spectra for 50% of full load; (a) 0.45% voltage unbalance, (b) 10.6% voltage unbalance

induction motor. Different levels of supply voltage imbalances were introduced by connecting a variable resistor in series with one of the motor phases. For each level of voltage imbalance the leakage flux spectrum were recorded. This was done for 25%, 50%, 75% and 100% of full load. Selected results are presented in Fig. 1 and Table 2.

Table. 2. Harmonic frequencies of leakage flux spectrum due to voltage unbalance at 50% of full load

Voltage imbal.	Harmonic Frequency	Amplitude	Change in Amplitude
%	Hz	dB	dB
0.45	98.51	-24.8	
0.45	150.02	-21.9	
6.88	98.26	-7.5	17.3
6.88	149.94	-10.5	11.4
10.60	98.69	4.3	29.1
10.60	149.93	-8.9	13

The flux spectra showed significant increase in the (2-s) harmonic. This confirmed the previously published findings [1, 3]. What is also clearly visible is the increase in 3 x supply frequency harmonic. This is, however, just an observation not supported by theoretical findings.

3 SHORTED TURNS AND PHASE TO PHASE SHORTS

Early detection of interturn shorts during motor operation would eliminate consequential damage to the adjacent coils and the stator core reducing repair costs and motor outage times. This fault is caused by the destruction of the insulation. According to researchers in [2], shorted turn fault can migrate to phase-to-phase or phase-to-ground shorts.

3.1. Predicted axial flux harmonic frequencies due to shorted turns and phase-to-phase shorts

In [4] researchers presented a complete theoretical and practical study of the evolution of axial flux spectrum. They showed that when the stator winding of an induction motor has any shorted turn the leakage flux harmonics of the following frequencies increase in amplitude:

$$f = \left(k \pm n \cdot \frac{(1-s)}{p} \right) \cdot f_s \quad (4)$$

where k = integer, 1,3 and $n = (2p - 1)$

Except from the frequencies in equation 4, changes

were detected in the electric motors which indicates turn-to-turn shorts by looking at the family of slot pass frequencies. These frequencies occur at Principle slot pass frequency (PSP) plus or minus $(2 \times N \times LF)$. And PSP frequency occurs at either the product of number of rotor bars (RB) and running speed, minus line frequency or product of the number of stator slots (SS) and running speed minus line frequency [5]. This is summarised in the equation below.

$$f = PSP \pm 2N \times LF \quad (5)$$

Where f = family of slot pass frequency, Hz
PSP = principle slot pass frequency, Hz, $Rb \times RPS - LF$
 N = any integer, 0,1,2,3....
 LF = line frequency, Hz

Table 3: Predicted flux harmonic frequencies due to turn-to-turn fault

Supply Frequency Hz	Rotor speed RPS	Harmonic Frequency Hz
50	24.32	25.68
50	24.32	74.32
50	24.32	98.64
50	24.32	122.96
50	24.32	147.28

Table 4: Harmonic frequencies of leakage flux spectrum due to turn-to-turn shorts at 50% of full load

Harmonic Frequency Hz	Amplitude before fault dB	Amplitude after fault dB	Change in Amplitude dB
24.68	-28.9	-13.5	15.4
74.32	-30.4	-13.4	17.0
98.64	-27.7	-25.3	2.4
122.96	-35.0	-21.3	13.7
147.28	-20.1	-15.4	4.7

Interturn fault was introduced by removing insulation of about 20 mm length of two neighbouring turns of a phase. These two turn were soldered together. The phase-to-phase fault was introduced by removing insulation and soldering two of the phases. The leakage flux spectra were captured for 25%, 50%, 75% and 100% loads. Selected results are presented in Table 4 and Fig. 2.

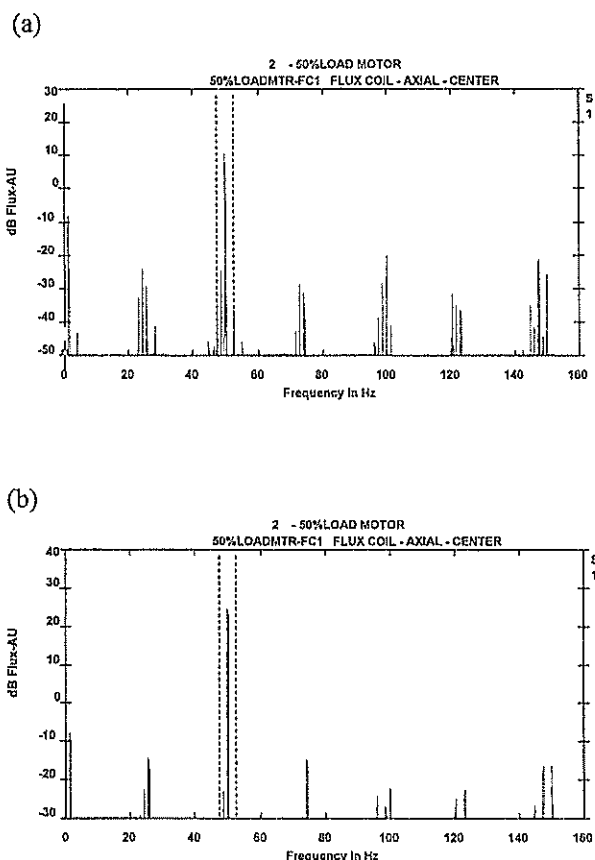


Fig. 2. Leakage flux spectra for 50% of full load; (a) motor in good condition, (b) motor with turn-turn short

The flux spectra showed significant changes in amplitude of most of the harmonics predicted in Table 2. The changes were higher for the phase-to-phase shorts. The changes in amplitude of this harmonics were observed for all load levels.

CONCLUSIONS

The flux coil is a very useful tool for detection of faults in induction motors. It is fully non-invasive technique, the coil is also very easy to mount at the back of the motor.

The leakage flux method proved to be specifically useful for the detection of the supply voltage unbalance and stator winding shorts. This method can be also used for the detection of other faults like airgap eccentricity and broken rotor bars but further research is required in these two areas.

ACKNOWLEDGEMENTS

The authors would like to thank Anderson and Hurley Instruments (S.A.) for lending hardware and software which was used to capture and analyse data.

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A DESIGN TOOL TO FACILITATE A MATLAB/ SIMULINK SIMULATION, TO RUN IN REAL-TIME OVER A NETWORKED DSP CARD

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ABSTRACT: *This paper describes the development and application of a tool, that allows Simulink to be run on a remote Digital Signal Processing card. Emphasis is placed on the technologies involved and the numerous possibilities that this tool will make available to both academic and commercial sectors.*

Keywords: Real-Time, Code generation, Windows Sockets, Simulink

1. INTRODUCTION

This project is an extension of a stand-alone system which allows Simulink® to run, in real time, on a PC32 Innovative Integration DSP card [3]. This tool enables one, to prototype controller with a minimal amount of time, Fig.1 illustrate this procedure.

Description of blocks in Fig. 1

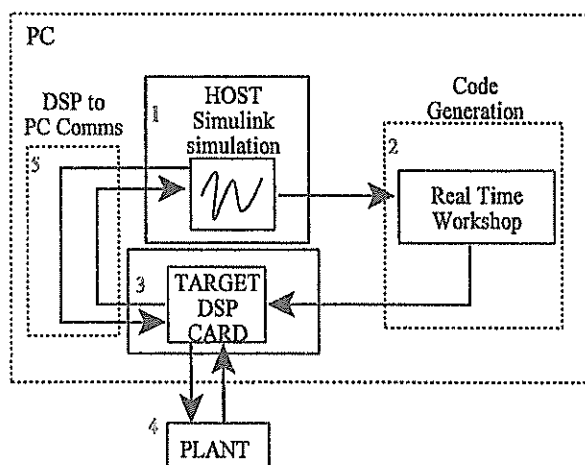


Fig. 1 *Stand-alone development tool*

1. This is a simulink simulation that is developed on the PC.
2. The Real-Time Workshop is an add-on toolbox provided for use with simulink. It converts a simulation into real-time code. A more detailed discussion is covered in section 3.
3. The DSP card resides in the PC and interfaces to it via an ISA slot.
4. The plant can be any system , and interfaces to the DSP card by either analog or digital I/O.
5. This block represents the communication channel that exists between the DSP card and the PC. The channel is implemented using a dual port ram technique.

This paper describes how the system in Fig.1, is to be expanded to run over a network. This means that the DSP card will be, located in a PC, at some remote site and

multiple user can gain access to it.

2. The Concept

To put the project into perspective, consider the following scenario, where a controller has to be developed and run on a remote DSP terminal. The development tool must allow the development terminal and DSP hardware to be link by a network, and process data must also be returned to the development terminal as detailed in Fig. 2.

To implement the system in Fig. 2, using this tool will require the following:

1. A development terminal running Matlab and Simulink.
2. A remote PC and DSP card interfaced to the plant.
3. Develop and simulate the desired controller.
4. Download and test on the remote DSP terminal.
5. If not satisfactory go to step 3.

That is how simple, rapid prototyping controller development becomes using this tool.

3. Matlab and Simulink

Matlab and Simulink are becoming widely used in academic and even industrial environments, and is therefore the natural choice for a simulation package. The simulation environment or Simulink allows automatic code generation via the Real-Time Workshop(RTW). The RTW allows ANSI C code to be generated from a Simulink diagram, Fig 3 details this process [2]. The RTW also handles the compilation and downloading of the code. The whole process is transparent to the user.

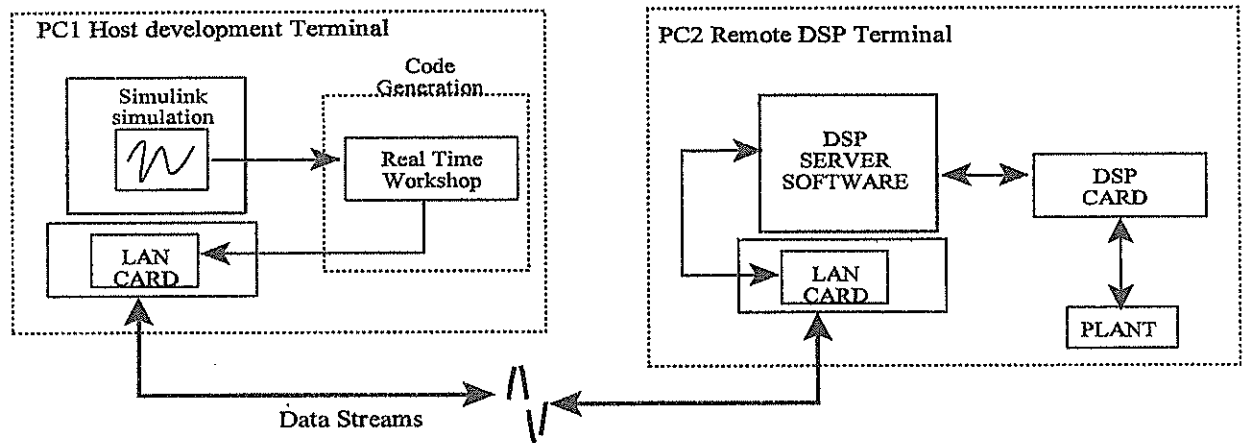


Fig. 2 Network Enabled Development Tool

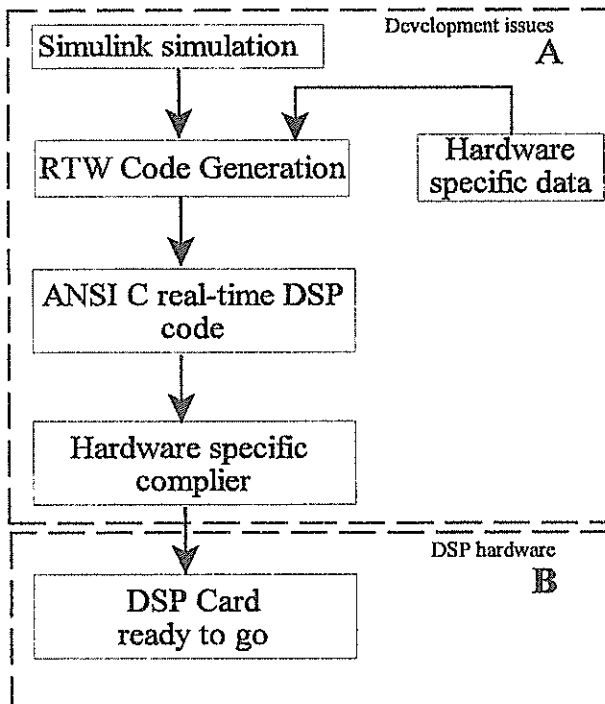


Fig. 3 Real Time Workshop Overview

4. Networking the system

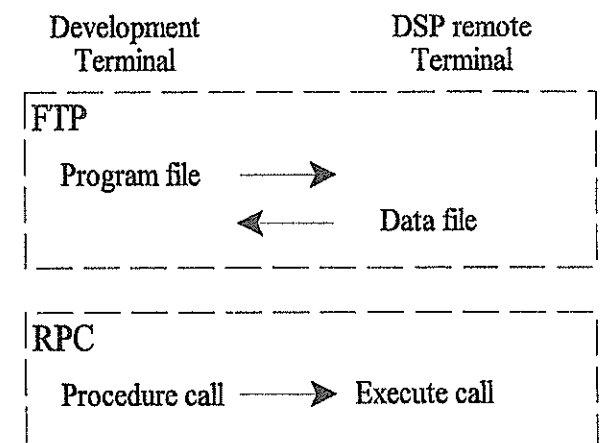
Fig. 2 shows a networked version of the system in Fig 1, with the help of Fig. 3, it is possible to insert a network between the development aspect and the DSP hardware. Block A in Fig. 3 contains the development side, while block B contains the DSP hardware. It is now possible to quantify the types of data streams, that can flow between the development and DSP terminals, as shown in Fig 2. For the requirement specified in section 2, only two types of data streams are necessary for the required functionality, this is shown in Fig 4.

Type 1: A pseudo File Transfer Protocol(FTP) data stream, which is used for program downloading and off-line data logging

Type 2: A Remote Procedure Call (RPC) protocol allows "meaningful" commands to be exchanged between two systems, with each system interpreting them and performing an appropriate action. For example, to start the code running on the DSP station, the development terminal will issue a START PROGRAM command and DSP station will respond accordingly.

Having defined these two data streams, a question which now arises is how can one implement the network software. Firstly one can go right to the communication stack or secondly one has the option to use a Application Program Interface (API). An API abstracts the network to a much higher level and makes programming more standardized, and therefore improves "portability" across the platform the system is running on. Since Microsoft Windows is the platform our system is running on, Windows Sockets is the natural choice.

Fig. 4: Data flow types



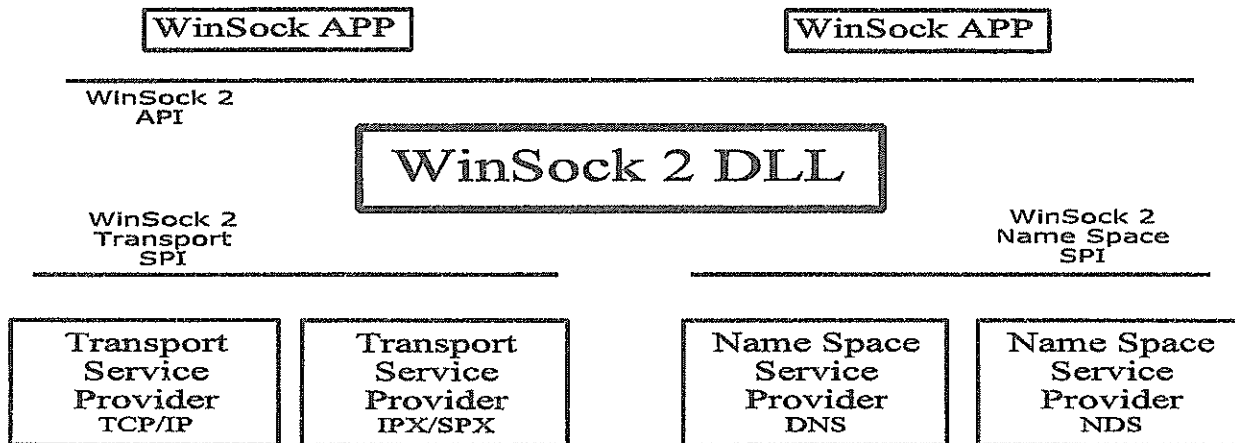


Figure 5: Windows Socket Architecture

5. Socket programming and Windows Sockets 2

Socket programming abstracts a network to communication end-points, that is, data is sent through a socket and appears on the other end of the socket. The socket programming paradigm was popularized by the Berkeley software distribution, which is for a UNIX system with a TCP/IP stack[4].

Windows Socket version 1 (WinSock 1) implements this concept on the Microsoft Windows platform. WinSock 2 extends the functionality of WinSock 1 programming by allowing multiple service providers to provide support for their communication stack via the common WinSock API. This is accomplished by having a Service Provider Interface (SPI) and a API. Fig. 5 details the relationship between the API and SPI [4].

6 Application

When designing this development tool, the concept of modularization was used to allow later enhancements and improvements to be easily incorporated. Since this system is a development tool, it is left to the designers' ingenuity and imagination to utilize its power and flexibility. Three possible applications are outlined below.

6.1 Educational Environment

In the educational environment the teaching of controller design is difficult and time consuming. Analog hardware has limitations while DSP hardware is costly and complicated to use. With this tool a user only has to produce a Simulink model of the controller and then download it to the DSP server. This system minimizes costs, since one DSP card can serve multiple users. The user will also have the flexibility of changing controller

parameter and seeing immediate results. Further more dangerous plants with moving parts can be isolated from the user, and prevent accidents. Fig.6 details this system.

6.2 Producing Advanced Application Specific Controller

The traditional approach involves simulation and embedding the controller the latter being the most time consuming. With this tool the embedding stage is short circuited. This results in a major time and cost saving. Due to the modular design of the tool tweaking and controller performance is both feasible and easy. Since network support is also included and implemented in a modular fashion, it may be easily incorporated into the final design irrespective of the network protocol used.

6.3 Data Acquisition applications with pre-filtering

In numerous data acquisition applications an additional amount of filtering or "data selection" is needed to access the useful data. This development tool will allow for this scenario, in addition on-line and off-line data logging will also be available. This tool will afford user the opportunity to develop specialized data acquisition cards, pertinent to their applications.

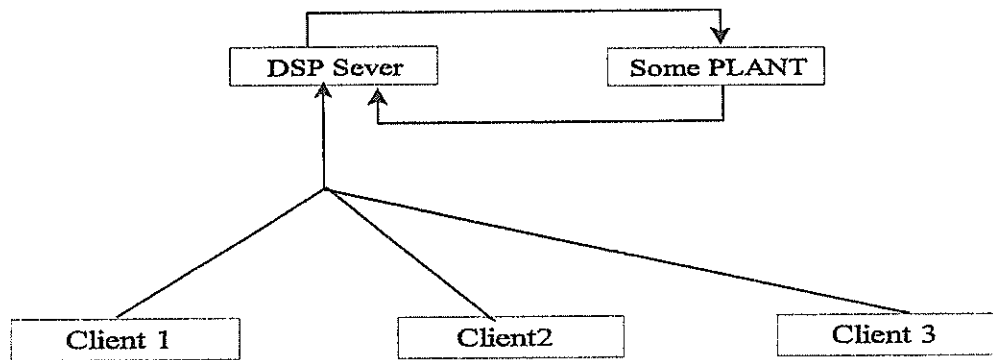


Fig. 6: Multiple Client Model

7 Conclusion

This system has a number of advantages and is a fraction of the cost of similar development tools. A great advantage of this tool is that users from any scientific or technical background with Simulink experience can make use of advanced DSP, without having to go through the messy DSP architecture and code.

The concept of automatic code generation, though no new, will enable faster design time and improve the time to market cycle, or in the case of researchers time to conference and time to MSc /PhD cycle. With processing power becoming cheaper by the day, high level design tools will open a new realm to developers and researches, this tool will hopefully allow South Africa to also be apart of this technology.

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ANALYSIS OF RADIAL MAGNETIC BEARINGS

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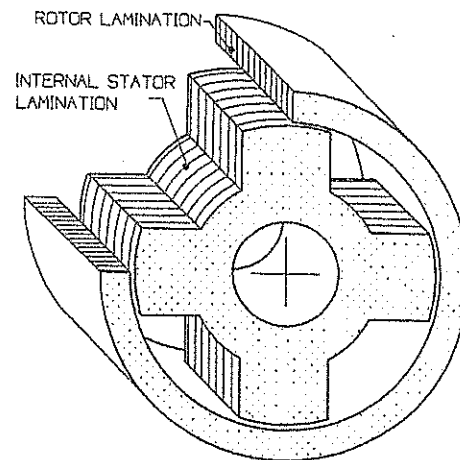
Abstract: A homopolar biased radial magnetic bearing is described. The magnetic field within the bearing was analysed using both 2-D and 3-D finite element models. The results of these analyses were used to calculate various items of the bearing performance: the load capacity, iron loss and bearing constants.

Keywords: magnetic bearings, electromagnetic actuators, magnetic field analysis, finite element analysis

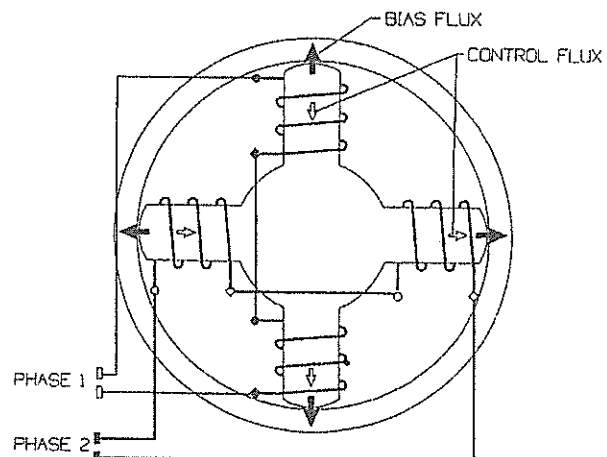
1 INTRODUCTION

A magnetic bearing system can be advantageous for rotating machinery due to the absence of mechanical wear, elimination of lubricating oil, and ability to operate at higher speeds than mechanical bearings. The forces required to support the rotating shaft are developed by an electromagnetic actuator, typically a part of a feedback control system. The electromagnetic actuator of a magnetic bearing has a rotating member (or rotor) and a stationary member (or stator) concentrically located with respect to each other. The rotor may be located concentrically internal or external to the stator. A magnetic bearing uses adjustable electromagnetic force generated by a current flowing through coils wrapped around the stator poles and controlled by a control circuit, to adjust the distance between the stator and rotor. Also, proximity sensors are typically used to measure the length of the gap and to provide input signals to the control circuit. In addition to the variable electromagnetic control flux, a magnetic bearing may also have a constant DC bias flux, which premagnetizes the working air gaps to a specified level of the magnetic flux density. The bias linearises the control laws of the bearing. In some configurations, with a homopolar bias, a permanent magnet can be used to provide the bias flux, eliminating the ohmic losses associated with the bias coils. Magnetic drag and iron losses are created whenever the rotor rotates within a spatially varying bias magnetic flux. Certain high-speed applications are very sensitive to rotor heating caused by core losses in the bearing rotor, and to the magnetic drag caused by losses in the rotor iron. The higher the speed the greater the losses, and increased heat can demagnetise permanent magnets which may be embedded in the rotor. Furthermore, if the rotor is located in a vacuum containment vessel it may be difficult to cool the spinning mass. A magnetic bearing design with very low rotor loss would be advantageous for such critical applications.

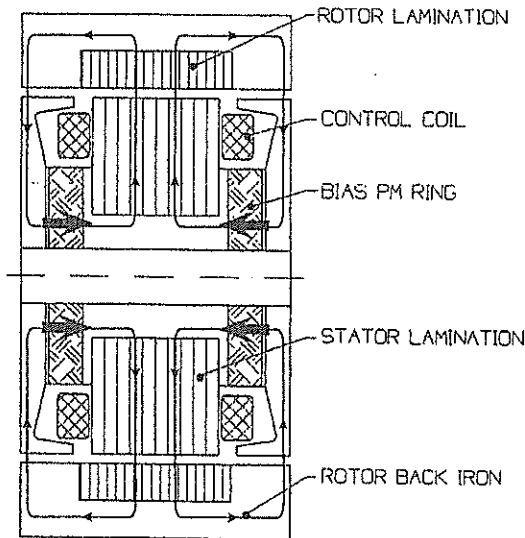
(a)



(b)



(c)



(d)

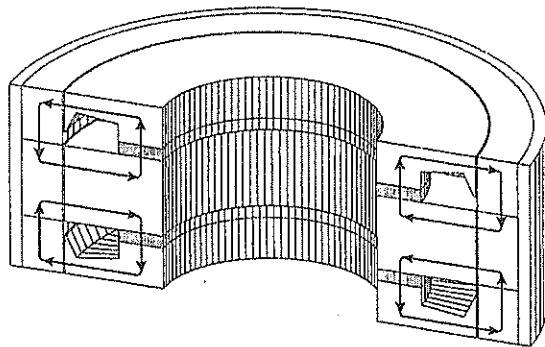


Fig. 1. Permanent magnet biased radial magnetic bearing with external rotor: (a) simplified general layout, (b) cross-section showing control flux path for the case when phase 1 carries control current, (c) longitudinal section showing bias flux path, (d) 3-D view showing bias flux path

Novel homopolar magnetic bearing concepts were described and patented by Malsky [1] and Meeks [2]. These concepts employ a homopolar flux to bias the stator poles to the same polarity, e.g. all are 'north' poles. Control windings direct the flux at opposite poles to one side of the stator or the other to produce a control force in the axis of the poles. Fig. 1 shows the configuration of the homopolar bearing with four poles, described in the patent [1]. The rotor lamination is subjected to a time varying magnetic flux as it rotates through the spatially varying time-invariant bias

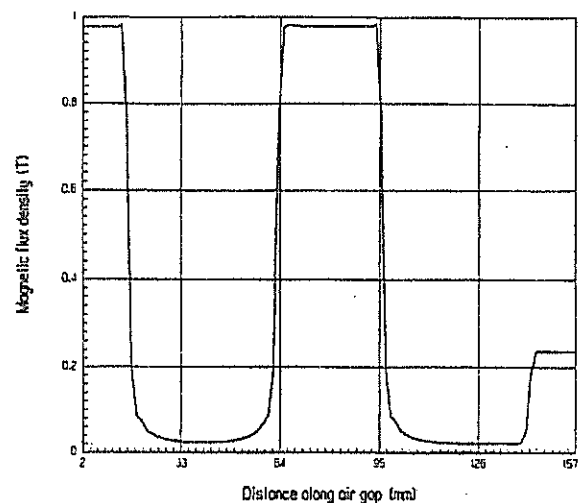
flux. The perceived advantage of the homopolar bearing is that the bias flux is of one polarity, hence the variation in the rotor flux will be reduced, and the rotor losses should be reduced relative to a bipolar magnetic bearing. Despite this feature, the resulting eddy current and hysteresis rotor losses, caused by discontinuities or sharp changes in the flux distribution along the air gap circumference (Fig. 2 (a) and (b)) can be unacceptable in certain applications, where an extremely low magnetic drag is required or where rotor heating is critical.

2 MAGNETIC FIELD ANALYSIS

To design 3-D finite element (FE) model which fully reflexes a complex 3-D distribution of both bias and control magnetic fluxes, as well as to perform magnetic field analysis is a very time consuming task. For the designer of a magnetic bearing the most interesting parameter is the magnetic force acting on the rotor. Since this force is created mostly in the air gap between the stator and rotor, the FE model can be reduced to two dimensions. To account for the bias magnetic flux, which is generated by permanent magnets located in the "third dimension", specific boundary conditions are required [3].

The finite element analysis of the described magnetic bearing was performed. The main objective of this analysis was to calculate the magnetic field distribution. The results of this analysis were in turn used to calculate the rotor iron loss, the force acting on the rotor (load capacity) and the bearing constants, as described in the following sections. The finite element analysis also provided interesting results which explain

(a)



(b)

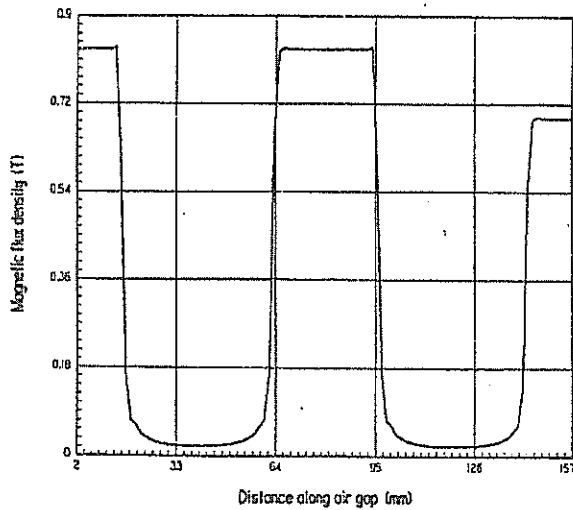


Fig. 2. Magnetic flux distribution along the air gap between the stator and rotor (half of the circumference plotted): (a) at full control current, (b) at 20 % of control current

a relatively high level of the iron loss in the bearing rotor. Fig. 2 (a) and (b) show the magnetic flux density distribution in the air gap of the bearing for two different values of the control current. It is clearly visible that the magnetic flux density changes rapidly from its maximum value, at the stator pole area, to practically zero, at the stator interpolar area. One should also notice four periods of the magnetic flux density change (four maxima and four minima) when travelling along the stator of the bearing. All these contributes towards the rotor iron loss of the bearing.

2.1 Results

The method introduced in this section was applied to the load capacity (force) calculation of a high speed (35 000 rpm) radial, homopolar magnetic bearing with a permanent magnet bias [3]. Both 2-D and 3-D finite element (FE) models of the bearing were built and the results of the load capacity calculation with the help of these models were verified experimentally. In both cases the virtual work method of force calculation, based on the magnetic field solution, was used [4, 5]. The values obtained from the FE analysis were compared with the force measured on the prototype bearing. As can be seen from Table 1, the agreement between calculated and measured values is very good (4 %).

Table 1: Comparison of calculated and measured force acting on the rotor; control winding current 10 A, rotor offset = 0

Force [N]		
Calculated		Measured
2-D FE model	3-D FE model	-
512	523	534

3 BEARING CONSTANTS

The finite element method was also used to determine the magnetic bearing constants.

The magnetic bearing force acting on the rotor when it is moved off-centre in, say, the x-direction is given by the following equation:

$$F_x = K_{bc} i_c + K_{bx} \Delta x + K_{cx} i_c^2 \Delta x \quad (1)$$

where:

- K_{bx} = bias field constant
- K_{cx} = control winding constant
- K_{bc} = mutual constant for bias field and control winding
- i_c = control current
- Δx = rotor displacement in x-direction

The values of all three constants are given in table 2.

Table 2: Calculated bearing constants

B_r	K_{bc}	K_{bx}	K_{cx}
T	N/A	N/mm	N/mm-A ²
0.8	51.2	1615	6.97

4 ROTOR IRON LOSS CALCULATION

The iron loss calculation was performed only for the rotor lamination. The solid rotor backiron (the outside shell shown in Fig. 1 (d)) conducts the uniformly distributed constant bias flux. Moreover, if any changes in the magnetic field distribution on the surface of rotor solid ring occur, the skin effect reduces the affected area to a negligible depth, (for example, the magnetic field equivalent penetration depth for

4340 Steel at frequency 583 Hz - 35,000 RPM is about 0.7 mm). In conclusion: it was assumed that the solid backiron of the rotor does not produce significant power losses due to the magnetic field variation.

For loss analysis the rotor lamination was divided into a few layers (concentric cylinders in the radial direction). The finite element analysis provided distribution of the magnetic field in each layer of the rotor. The field distribution in each layer was different and depended on the level and direction of the external load. The radial and tangential components of the magnetic field density were separately analysed using the Fast Fourier Transform. The spectrum of harmonics was used to calculate the hysteresis and eddy current losses. The harmonics magnitudes of field spectrum lower than 0.025 T were neglected to simplify the calculation. The rotor losses computed for two different rotational speed are presented in Table 3 [6].

Table 3: Comparison of rotor losses at 20% load (control current 2 A), and at full load (control current 10 A).

Rotor Losses (W)			
Rotor Speed 17,000 RPM		Rotor Speed 35,000 RPM	
20% Load	100% Load	20% Load	100% Load
25.3	36.1	64.5	92.9

CONCLUSIONS

A homopolar radial magnetic bearing with a permanent magnet bias has been described.

Finite element method was used for the magnetic field analysis in the described bearing. The method proved to be a perfect tool for analysis and performance calculation of devices having a complex 3-D magnetic flux distribution.

For some calculations, e.g. the load capacity or bearing constants, a very accurate but time and resources consuming 3-D finite element modelling can be replaced with a much faster and cheaper 2-D FE modelling. Despite its simplicity, the 2-D approach offers good accuracy.

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RIPPLE FREE BRUSHLESS LINEAR SERVO MOTOR

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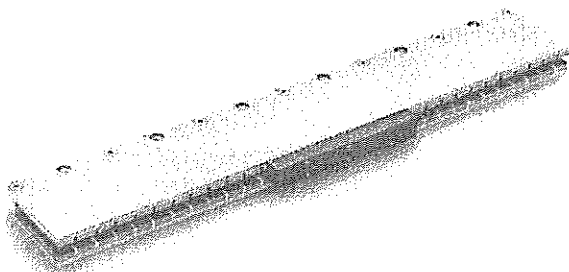
Abstract: This paper deals with a permanent magnet linear brushless motor, which demonstrates practically ripple free thrust. Two and three-dimensional finite element models were used to calculate thrust and ripple. Comparison of FEM results with test data has shown a good agreement.

Keywords: linear permanent magnet motors, brushless linear motors, finite element magnetic field analysis

1 INTRODUCTION

Permanent Magnet brushless motors can be classified into rotary motors and linear motors. Due to advances in power semiconductor technology interest in these motors has grown over the last few years [1, 2]. Double Sided Brushless Linear Servo Motors (LSM) are available commercially for more than 10 years. The motor construction is shown in Fig 1. The stator (primary part) consists of not overlapping coils placed side by side in the epoxy core. The secondary part contains two parallel steel plates with NdFeB permanent magnets mounted on the internal surfaces. The armature (stator) is placed in the air gap between the two permanent magnet assemblies. The main advantages of the motor are: (1) no net normal force between primary and secondary parts, and (2) a ripple free thrust at reasonable thrust-to-volume ratio. This was achieved by using a coreless armature, high energy permanent magnets and through optimisation of magnet pitch to armature coil-span ratio.

(a)



(b)

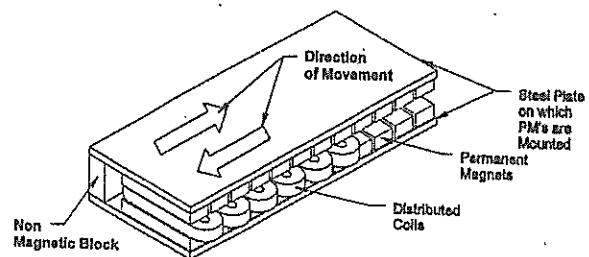


Fig. 1. Balanced Brushless Linear Servo Motor: (a) general view, (b) components

The motor is applied in versatile motion systems ranging from straight units to multiple X-Y-Z configurations. The brushless balanced LSM is available commercially in various models with standard travel from 50 mm to 3000 mm. Basically, there is no limitation on motor travel displacement, since the stationary parts can be easily joined together to form any length.

Linear motors generally provide the smoothest motion. Particularly, brushless linear servo motors provide extremely smooth motion having no contacting surfaces causing jitters. The epoxy core motors have been optimised to provide minimum velocity ripple. Advanced magnet design, non-ferrous epoxy core, sinusoidal commutation, and linear servo amplifiers are just a few of the technological advantages enabling achievement of velocity ripple on the order of 0.01%. By contrast, a ball screw arrangement is not as smooth due to the vibrating nature of balls entering and exiting the nut raceways. This is easily observed in sub-micron systems. Belt and rack-and-pinion drives also have contacting mechanisms which are susceptible to friction and backlash caused vibrations.

Linear motors are essential components for achieving very high position accuracy. They are employed in systems designed for optimal performance on sub-micron scale. Typically, a brushless linear servo motor is used with a linear encoder or laser interferometer transducer. Position can be controlled from 0.025 mm to 0.000025 mm accuracy dependant upon the feedback device.

Balanced linear motors (with zero preload) are the most suitable devices for application in the systems with high dynamic requirements. They are capable of very high velocities (10 m/s) and accelerations (10g). The absence of mechanical windup, backlash and friction in the drive often results in setting time of a few milliseconds in a carefully designed system. Ironless core linear motor system demonstrates close loop position bandwidths of 400 Hz with capability of producing 2500 N peak force.

The thrust ripple investigation was performed on Brushless Linear Motor type LEB-S-2-S [3] with no cooling system.

2 MAGNETIC FIELD ANALYSIS

Finite element analysis has been performed to calculate magnetic field distribution and to facilitate calculation of thrust and the winding inductance [4]. Three dimensional finite element models were used, since they offer the most accurate representation of the motor and lead to the most accurate results.

The magnetic field solver computed the magnetic field using Ampere's law:

$$\nabla \times \mathbf{H} = \mathbf{J} \quad (1)$$

and Maxwell's equation describing continuity of the magnetic flux:

$$\nabla \cdot \mathbf{B} = 0 \quad (2)$$

where:

$\mathbf{H}(x,y,z)$ is the magnetic field,

$\mathbf{J}(x,y,z)$ is the current density

The magnetic flux density $\mathbf{B}$ was computed using the relationship:

$$\mathbf{B} = \mu \mathbf{H} \quad (3)$$

where $\mu(x,y,z)$ is the permeability.

Selected results of the finite element analysis are presented in Fig. 2 to Fig. 5. As can be seen from Fig. 2 the magnetic flux density distribution in the air gap is very regular, sinusoidal. The magnitude is around 0.5 T which is reasonable result for a motor with coreless stator.

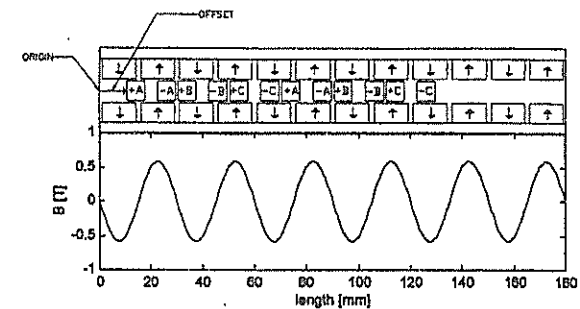


Fig. 2. Magnetic flux density distribution (calculated) in the middle of the air gap between two PM assemblies.

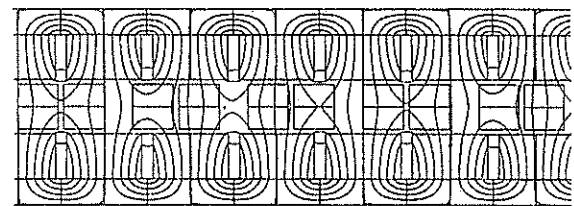


Fig. 3. Magnetic flux lines in the cross-section of the motor.

The only elements of the motor which could be subjected to saturation are steel plates of the PM assemblies. Fig. 4 shows the magnetic flux density distribution in the cross-sectional area of the motor, for the maximum possible value of the stator current. The highest values of the flux density in plates appear close to the gaps between magnets but the flux density is still below the saturation level.

Since the air gap is the most important part of the motor (this is where thrust is produced) the magnetic flux density distribution in this area was used to verify the accuracy of the magnetic field calculations. The results presented in Fig. 5 show a good agreement between calculated and measured values.

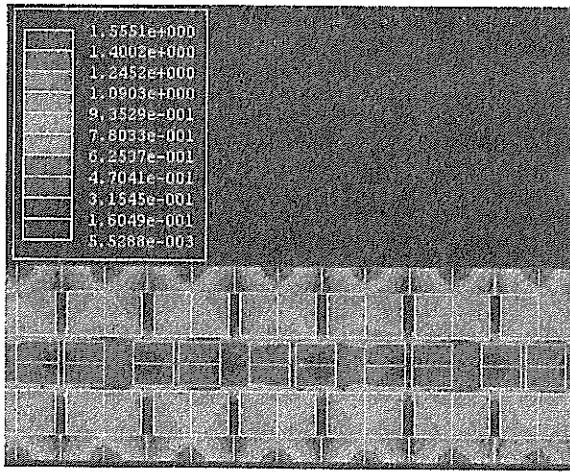


Fig. 4. Magnetic flux density distribution; currents in the winding were adjusted to give the maximum thrust, $I_m = 9$ A.

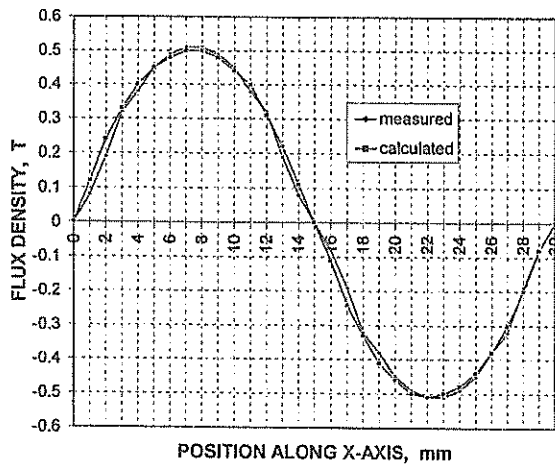


Fig. 5. Magnetic flux density distribution in the middle of the air gap between two PM assemblies and in the centre of the magnet length (two magnet pitches).

4 MOTOR PERFORMANCE ANALYSIS

4.1. Principle of Thrust Production

The motor under consideration is a three-phase permanent magnet synchronous motor. The two PM assemblies (Fig. 1) produce sinusoidal magnetic flux density distribution in the winding region (see Fig. 2). When a symmetrical 3-phase system of currents is applied to the winding a force between the winding and the PM assemblies is created. This force has only x-component. Due to symmetrical position of the winding with respect to the PM assemblies, y- and z-components of the force vanish. The x-component of

the force (thrust) depends both on the position of the winding with respect to the PM assemblies, and the phase shift of the currents in the winding. Therefore, the task of a drive supplying the winding is to keep adjusting the phase angles of the currents while the mover travels along the stator.

4.2. Thrust Calculation

Thrust calculation was based on a three dimensional finite element magnetic field analysis.

The component of the Lorentz force due to the current in a magnetic field was calculated as:

$$\mathbf{F} = \int_V \mathbf{J} \times \mathbf{B} dV \quad (4)$$

The objective of these calculations was twofold: (1) to verify the accuracy of finite element analysis, and (2) to find out whether the thrust at higher currents was influenced by armature reaction and saturation of permeable elements of the motor. The calculations were done for two different currents: (1) rated (3.8 A peak) and (2) maximum (9.0 A peak). The calculated values were verified by measurements. The results are summarised in Fig. 6 and Table 1.

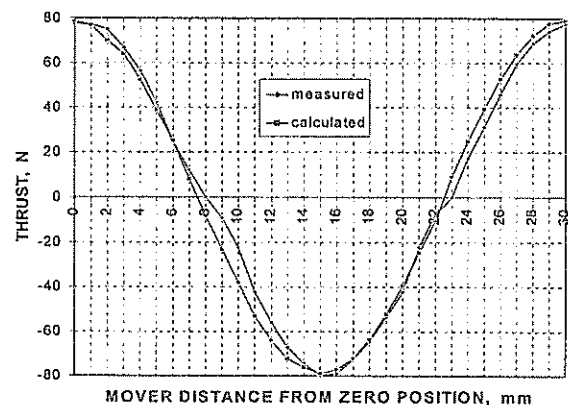


Fig. 6. Static thrust for different positions of the mover; $I_m = 3.8$ A

Table 1. Calculated and measured maximum values of the thrust

Phase current A (peak)	Thrust [N]	
	Calculated	Measured
3.8	78	80.3
9.0	184	182.7

The results presented in Fig. 6 demonstrate a good agreement between measured and calculated values of

thrust, proving usefulness of the presented method. The relationship between current and the maximum value of thrust is almost linear, meaning that even at a higher currents there is no visible armature reaction or saturation of the magnetic circuit.

4.3. Thrust Ripple Investigation

An inherent problem of a typical linear PM synchronous motor having winding located in slots is thrust ripple [5, 6]. The main cause of the thrust ripple is a variation of both the reluctance and magnetic flux density with the mover position due to armature reaction. In the motor described in this paper both problems were avoided by the introduction of the coreless stator design. One disadvantage of coreless motors is lower thrust to volume ratio as compared with motors having stator core. This problem can be minimised by a careful design. In order to investigate a possibility of the thrust fluctuation (ripple) the mover was shifted with respect to the stator over one magnet pitch, at 1 mm intervals, starting from the zero position. The phase angle of the currents was adjusted to give the maximum thrust for each position of the mover. The method of thrust calculation was the same as described in the previous section. The results of this investigation are presented in Table 2.

Table 2. Thrust vs. mover position; stator currents adjusted to give maximum thrust at each position, $I_m = 3.8$ A.

Mover distance from zero position	Thrust
mm	N
0	82.57
1	82.56
2	82.55
3	82.37
4	82.33
5	82.49
6	82.56
7	82.51
8	82.59
9	82.55
10	82.53
11	82.34
12	82.53
13	82.46
14	82.31
15	82.47

The difference between the maximum and the minimum values compared to the average thrust was 0.3%. This should be considered as a result of a

limited accuracy of the modelling method used, rather than the thrust ripple.

Conclusions

A balanced permanent magnet linear synchronous motor, which develops practically ripple free thrust has been presented. Accurate method of a three dimensional magnetic field analysis has been described and experimentally verified. The method of the motor's performance calculation, based on three dimensional magnetic field analysis, has been successfully applied.

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ANALYSIS OF A SHUNT CAPACITOR COMPENSATED ADJUSTABLE SPEED INDUCTION MOTOR UNDER FAULT CONDITIONS

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Abstract: This paper investigates the performance of a pulse width modulated (PWM) inverter fed induction motor drive system under balanced and fault conditions; disconnection of one phase from the inverter to the motor, disconnection of one phase at input to the front-end converter, and failure of a rectifier/ inverter switching device. Study has also been carried out on the influence of phase and voltage imbalances as well as the effects of shunt compensation in the system. Results obtained, which are discussed in the paper gave further insights to the behaviour of the system and its protection under fault conditions.

1.0 Introduction

Power electronics controlled adjustable speed drives (ASDs) have become increasingly popular in industrial applications, mining industry, assembly lines and HVAC systems. They have the advantage of improving energy utilisation and system efficiency and enhancing precision process control. There, however, exists growing concern for their operation under unbalanced and fault conditions, and their consequent influence on the design of suitable protection systems and power quality [1,2].

Adequate protection is essential to safeguard the equipment and improve reliability. Designing a reliable protection system, however, requires proper information of the behaviour of the system to be protected under the conditions of concern. The effect of compensation in ASD is to improve wave-form quality by reducing harmonic components which are introduced in the supply network from the front-end converter, and to the motor from the PWM inverter. Harmonics deteriorate motor drive performance and cause heating, over-voltage and equipment malfunction in the distribution and transmission circuit and other consumer loads. It is believed that these effects are exacerbated under imbalance and fault conditions [3].

There are many published work relating to the analysis of PWM based converters and voltage source inverters feeding balanced induction motors[4]. The analysis of faulted induction motor inverter drive has also been addressed[1]. These however, are inexhaustive in their treatment as vital information useful for designing adequate protection system and analysing faulted drive system for power quality mitigation purposes are still lacking.

This paper set forth the analysis of unbalanced and faulted shunt compensated induction motor drive system. The behaviour of the system under different unbalance and fault conditions are presented.

2.0 Non linear model of converter motor system

The overall system consists of an ac/dc buck rectifier feeding a PWM inverter through the dc link filter as shown in Figure 1. Shunt capacitors are placed at both ends of the rectifier-inverter system for harmonics distortion mitigation. Equations governing the converter model can be written as

$$p I_a = 1/L_{in} [V_{an} - r_{in}I_a - V_{ca}] \quad (1)$$

$$p I_b = 1/L_{in} [V_{bn} - r_{in}I_b - V_{cb}] \quad (2)$$

$$p I_c = 1/L_{in} [V_{cn} - r_{in}I_c - V_{cc}] \quad (3)$$

$$p V_{ca} = 1/L_{in} [I_a - S_{aif}] \quad (4)$$

$$p V_{cb} = 1/L_{in} [I_b - S_{bif}] \quad (5)$$

$$p V_{cc} = 1/L_{in} [I_c - S_{cif}] \quad (6)$$

The supply phase voltages are V_{an} , V_{bn} and V_{cn} with phase currents I_a , I_b and I_c . The voltages across the capacitors on the input side are denoted by V_{ca} , V_{cb} , V_{cc} and currents flowing into the rectifier as I_{ar} , I_{br} and I_{cr} . The dc side of the converter are defined as

$$p I_f = \frac{S_a V_{ca} + S_b V_{cb} + S_c V_{cc}}{L_f} - \frac{V_{df}}{R_f L_f} - \left(\frac{r_s}{R_s} + r_r \right) \frac{I_f}{L_f} \quad (7)$$

$$p V_{cf} = \frac{I_f}{R_x C d} - \frac{V_{cf}}{R_x R_L C d} \quad (8)$$

$$R_x = 1 + \frac{r_c}{R_L} \quad (9)$$

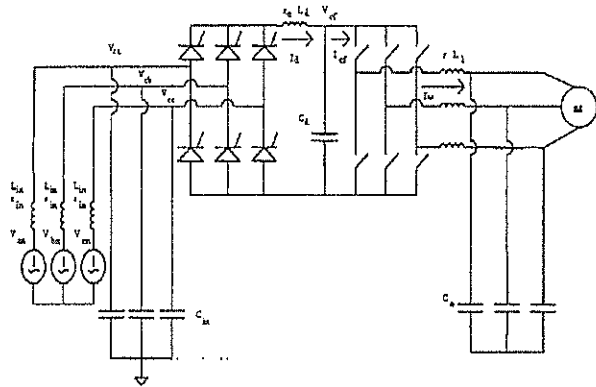


Figure 1. Shunt capacitor compensated PWM inverter fed induction motor.

where r_c is the resistance of the filter and R_L is the output of the filter viewed as a resistance load and $p = d/dt$. The three phase ac/dc buck converter is controlled by the sine-triangle pulse width modulation. The constraints on switching functions of the switches on the upper and lower half of the bridge are given in [4], as

$$S_{ap} + S_{bp} + S_{cp} = 1 \quad (10)$$

$$S_{an} + S_{bn} + S_{cn} = 1 \quad (11)$$

$$S_a = S_{ap} - S_{an}; S_b = S_{bp} - S_{bn}; S_c = S_{cp} - S_{cn}$$

2.1 Systems equations in stationary reference frame

The inverter motor drive system has been simplified and transformed in the d-q equivalent circuit as shown in Figure 2, with the derived equations given below. The currents flowing out of the PWM inverter denoted by I_{qs} and I_{ds} , while the voltages are given as V_{ds} and V_{qs} . The current flowing into the motor are I_{qsm} and I_{dsm} while the voltages appearing across the shunt capacitors are V_{qsm} and V_{dsm} .

$$I_{as} = I_{qs} \quad (12)$$

$$I_{bs} = \frac{1}{2} [-I_{sq} - \sqrt{3} I_{ds}] \quad (13)$$

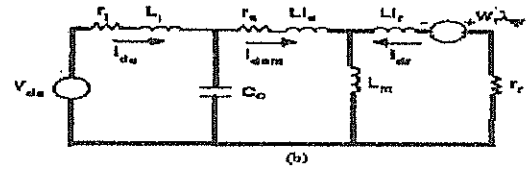
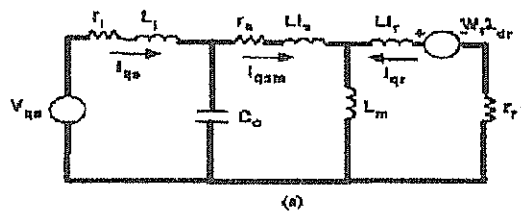


Figure 2. Equivalent circuit model of induction motor in (a) q-axis (b) d-axis.

$$I_{cs} = \frac{1}{2} [-I_q + \sqrt{3} I_{ds}] \quad (14)$$

$$V_{qs} = \frac{1}{3} [2 (S_a V_{cf}) - S_b (V_{cf}) - S_c (V_{cf})] \quad (15)$$

$$V_{ds} = -\frac{1}{\sqrt{3}} [S_b (V_{cf}) - S_c (V_{cf})] \quad (16)$$

$$V_{qs} = L_o p I_{qs} + r_o I_{qs} + V_{qsm} \quad (17)$$

$$V_{ds} = L_o p I_{ds} + r_o I_{ds} + V_{dsm} \quad (18)$$

$$p V_{qsm} = \frac{1}{C_o} [I_{qs} - I_{qsm}] \quad (19)$$

$$p V_{dsm} = \frac{1}{C_o} [I_{ds} - I_{dsm}] \quad (20)$$

$$V_{qsm} = r_o I_{qsm} + p \lambda_{qsm} \quad (21)$$

$$V_{dsm} = r_o I_{dsm} + p \lambda_{dsm} \quad (22)$$

$$V_{qr} = 0 = r_o I_{qr} + p \lambda_{qr} - \omega_r \lambda_{dr} \quad (23)$$

$$V_{dr} = 0 = r_o I_{dr} + p \lambda_{dr} - \omega_r \lambda_{qr} \quad (24)$$

$$p \omega_r = \frac{P}{2J} [T_e - T_L] \quad (25)$$

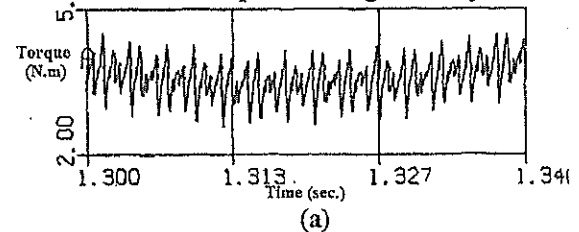
$$T_e = \frac{3}{2} \frac{P}{4} L_m [I_{qsm} I_{dr} - I_{dsm} I_{qr}] \quad (26)$$

T_e is the electromagnetic torque and T_L , the load torque. The electrical angular speed of the motor is given by ω_r .

3.0 Operation of the system

3.1 Balanced conditions

The motor drive system is simulated under normal operating conditions with and without shunt compensation. Instantaneous electromagnetic torque signal is shown in Figure 3(a), (b) for compensated and uncompensated conditions respectively. The mitigation effect of the shunt capacitor is significantly noted.



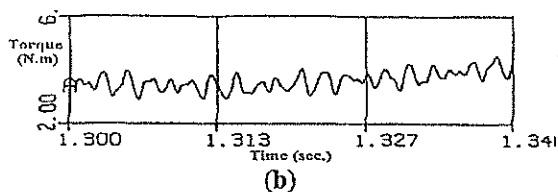


Fig 3. Electromagnetic torque in motor drive system.(a)compensated (b) uncompensated

3.2 Voltage and phase imbalances

A voltage imbalance of 25% was introduced in one phase at the supply. Simulated results showed the influence of such condition in the system. Under compensation, the rectified output voltage was less distorted than without compensation as shown in Figure 4 (a), (b). On the other hand, tests performed for unbalanced phase conditions (30°) resulted in a much distorted rectified waveform under both compensated and uncompensated conditions, in relation to the voltage magnitude as shown in Figures 5 and 6. This thus imply insignificant mitigation effects of shunt capacitor compensation under phase imbalance conditions. This case requires special converter topology or active filters to reduce harmonics distortion. The motor drive system being fed by the PWM inverter is dependent on the quality of the rectified voltage. Phase imbalances, therefore, will have detrimental effects on the operation of ASDs and other loads connected to the supply network.

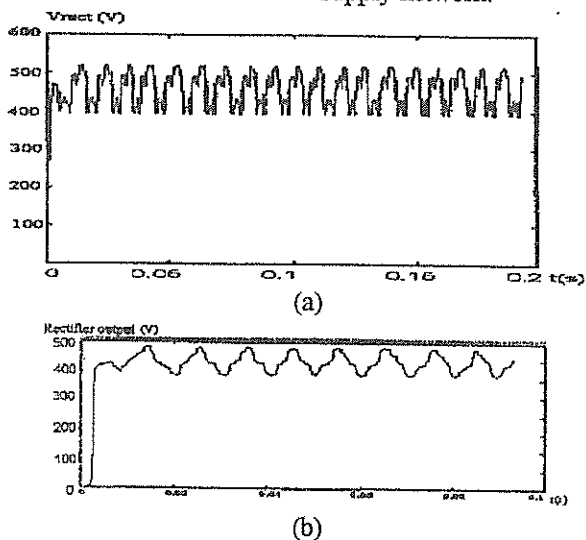


Fig 4. Effect under voltage imbalance on rectified voltage from dc link (a) uncompensated (b) compensated

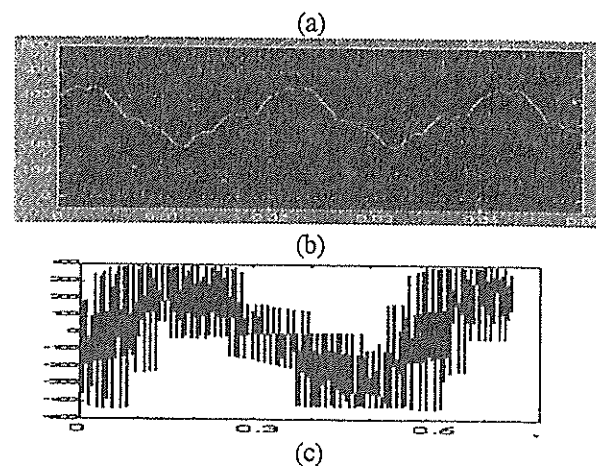
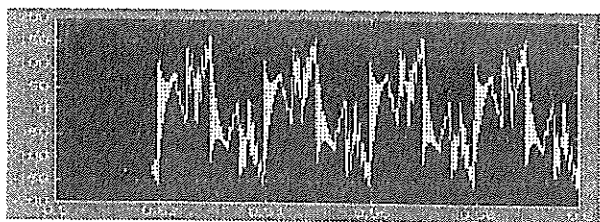


Fig 5 Phase Imbalances for uncompensated system (a) Input voltage (b) Rectified Output Voltage (c) Line voltage at Inverter Output

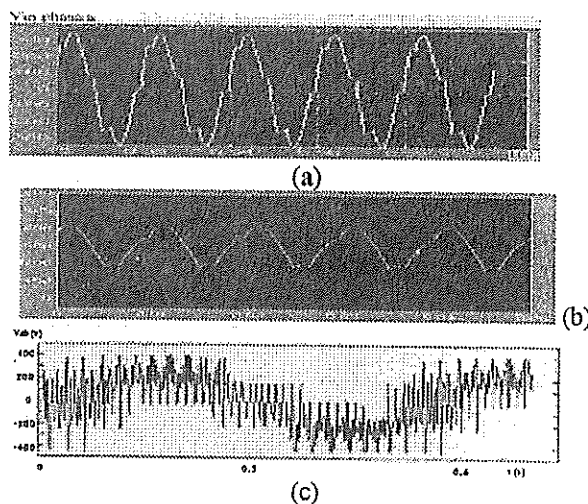


Fig 6. Phase imbalances under compensated conditions (a) Input Voltage (b) Rectified output voltage (c) Line voltage at Inverter output.

4.0 Faults conditions

4.1 Phase 'a' of the inverter open

With the inverter phase 'a' open a high degree of unbalance is introduced in the motor supply. This introduces input current imbalance with high current in phases 'b' and 'c', which may cause the damage of power electronic switching devices if protection devices do not operate. The high negative sequence current increases heating of the motor, lower efficiency and increase cogging torque of the drive motor. The resulting waveforms under this condition for compensated and uncompensated situations are as shown in Figures 7, 8 respectively.

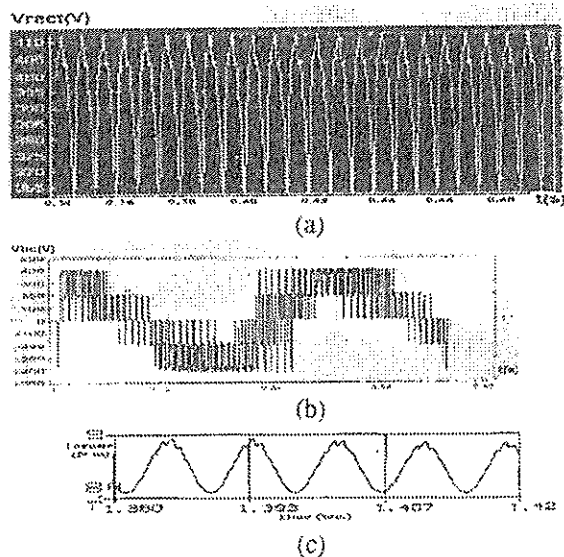


Fig 7. Inverter phase 'a' open, (a) Rectified voltage, (b) Inverter output voltage (c) Electromagnetic Torque (uncompensated system)

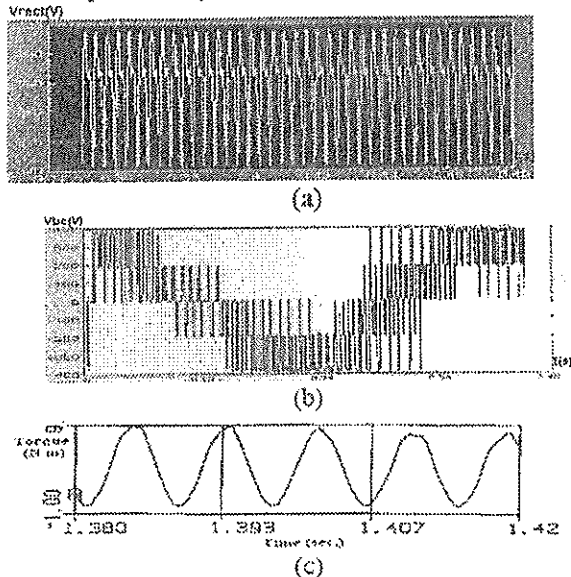


Fig 8. Inverter phase 'a' open (compensated system). (a) Rectified Output Voltage (b) Inverter output Voltage (c) Electromagnetic Torque

4.2 Failure of a switch in the converter

These faults are unpredictable and have pernicious effects on ASDs. [2] connected to the supply network. The simulated results obtained, Figure 9 show that the dc link voltage is badly affected well outside the expected voltage range, thus imposing higher requirements on the filter design criteria. Failure of switching device in inverter introduces asymmetrical operations.

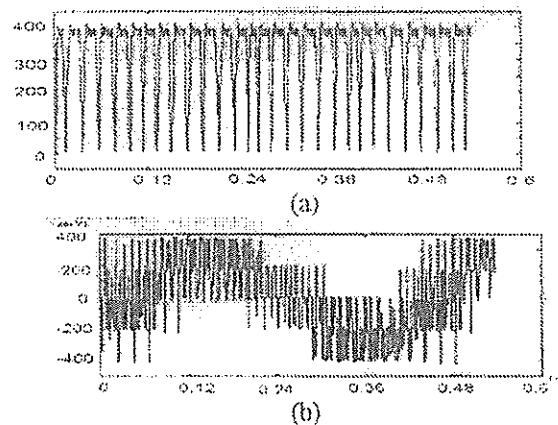


Fig 9.0 Failure of switching device (a) Rectified Voltage (b) Inverter output.

4.3 Phase 'a' of the supply open

This, like in section 4.1, also creates high degree of current imbalance which could be as high as 50 %. This could potentially overload the circuits causing an ASD trip condition [2]. The currents in the remaining two phases rise significantly. Meanwhile the waveforms, Figure 10, are distorted significantly increasing the level of harmonics distortion and asymmetrical operation.

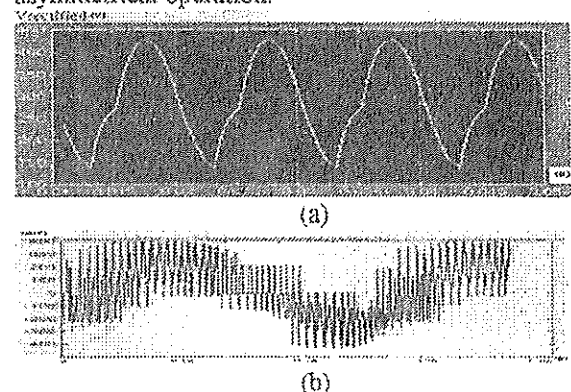


Fig 10. Input phase 'a' open (a) Rectified voltage (b) Inverter voltage feeding drive.

5.0 Conclusion

Effects of imbalances in the supply voltage magnitude and phase, and common fault conditions in a three phase PWM rectifier-inverter fed induction machine ASD with shunt capacitor compensation have been explored in this paper. It is shown that harmonics distortion of signals are exacerbated by most of these imbalance and fault conditions. The degree of asymmetry and distortion which ensue are quantifiable, and are important inputs to the design of the necessary protection system. The mitigating effect of shunt capacitors compensation are noticeable for most of the conditions investigated, except for the phase imbalance which requires other harmonic limiting means.

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SYSTEM INTEGRATION TECHNIQUE FOR POWER SYSTEM ANALYSIS

PA FOURIE, R. ZIVANOVIC

BAYSIDE ALUMINUM AND TECHNIKON PRETORIA

Abstract: In this paper we discuss the issue of integrating power system analysis tools in the hierarchical control structure of an industrial network. This will allow network operators to make quick and accurate studies by using real – time measurements. This will help the operators to understand the system limits and the effects of various switching actions. These tools will be also used for planning studies as well as training.

1. Introduction

System integration between Programmable Logic Controllers, Supervisory Control and Data Acquisition Systems, Relational Databases and Business Systems is a fast growing phenomenon currently in Industry all over the world. This integration of different systems has obvious advantages for businesses in terms of reliability, efficiency and productivity.

1.1 Plant Description

Bayside Aluminum in Richards Bay is one of ESKOM's biggest customers with a maximum demand of 350MW and together with its sister company, a demand of over a 1000MW.

The plant consists mainly of three potlines, each ± 1 Km long. Potroom A, the longest of the three, consists out of 240 pots and potroom B and C 148 and 140 respectively. The electrolyses process of producing aluminum uses a DC current of 143000 Ampere on the one potline and 116800-Ampere DC on the other two potlines respectively. The reduction process use requires the company to operate 24 hours a day 365 days a year.

The importance of a reliable and stable supply is absolutely crucial to the production of aluminum. These factors are almost directly proportional to the production of aluminum. They not only influence the control systems used on the pots but also predict the volume of aluminum produced each day.

Any outage to a potline will result in production losses that can never be made up again. An outage to any of the potlines for longer than six hours will have catastrophic consequences to both the company and its personnel, as it will result in a total plant shut down.

The pots will freeze up which will make it impossible to start up again. The vital importance and vulnerability of the system can not be overstressed. The following events illustrate this:

- December - 1973 - Flashover 33KV in comer B - No power for 3 hours 30 minutes.
- March 1977 - Feeder box to Rectifier (Rectifier Transformer Unit) A5 and capacitor CT blew up

during a high transient voltage condition - Rectifiers on potline A out for 3 hours 30 minutes.

- July 1978 - Rectifier diode bank blew up after failures of diodes.
- August 1996 - Contractor cut the wrong PT cable at a distribution station feeding Bayside Aluminum resulting in a 4 hour 30 minute outage.

1.2 Power system description

A dual redundant ring feed system ranging in voltage from 132kV, 33kV, 11kV and 400V. The plant is fed from four 195MVA transformers. Two on standby feeding the auxiliary loads and plant power and the other two feeding the 15 rectifiers of the three potrooms.

Harmonic filter banks filtering the 3rd, 5th, 7th and 11th harmonic generated by the rectifiers and correcting the plant power factor.

1.3 Control system description

Bayside Aluminum uses three possible ways of controlling the system local, common and central control.

The entire plant can be controlled and monitored from one central control room. All system I/O and data acquisition (measurements, circuit breaker status, control and metering) are connected to a Modicon Modbus Plus (protocol) PLC [1] (Programmable Logic Computer) fiber optic network.

Integrated to the PLC network is the next level of control. A Wonderware Intouch [2] SCADA (supervisory control and data acquisition) system. This is a highly versatile and powerful tool that is also integrated to relational databases or SQL servers' [2].

This type of control has become a standard in almost all big industries all over the world. Integrating power system analysis tools to this existing system is thus possible for any company who could see the advantages of such a system.

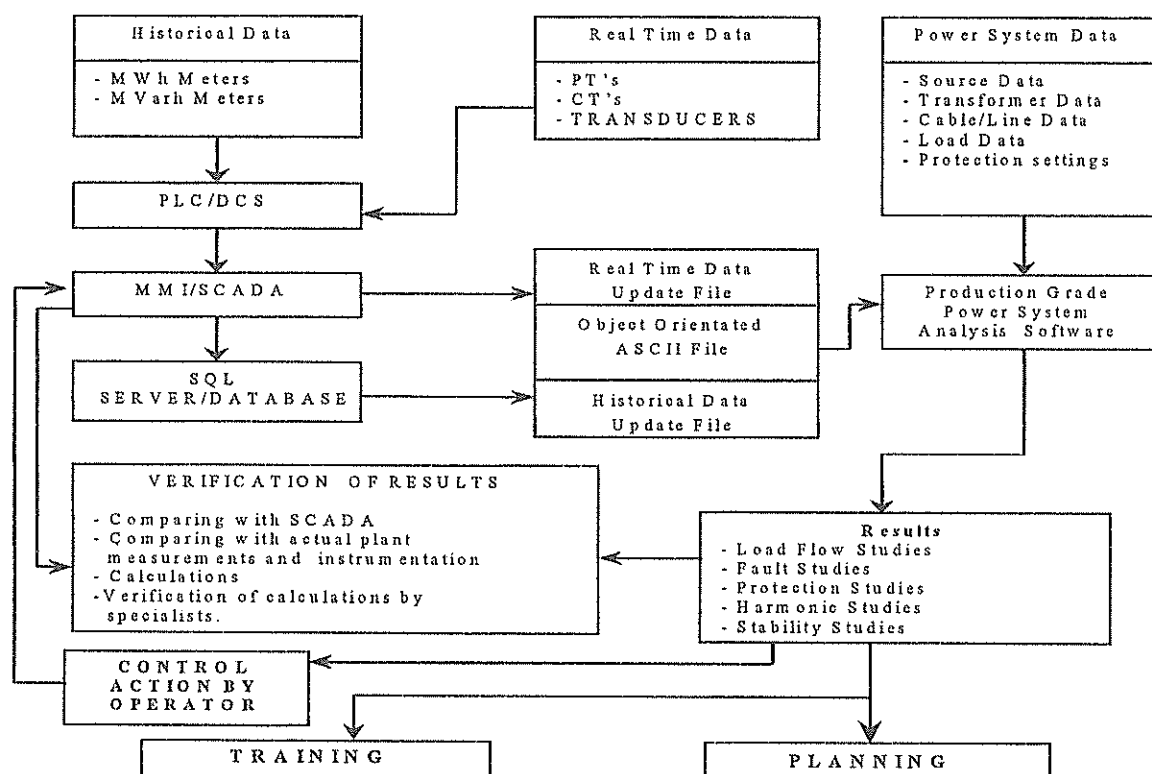


Figure 1: SYSTEM INTEGRATION DIAGRAM

2. Integrating computerized power system analysis

Three types of data are integrated into the PSAS (Power system analysis software), historical system measurements, real time system measurements and power system parameters.

Historical data is typically data obtained from the energy and metering system that is stored in a relational database. This is very accurate data of the system loading and consumption in a 24-hour period.

This type of data will normally be used with the aid of PSAS in feasibility studies to do system planning, modifications and extensions. It can also be used in personnel training to give them a good understanding of various switching methods and system limits.

Real time system measurements allow the user to take snap shots of the system in any instance of time to do various types of system studies. This instantaneous measurement can be updated in the PSAS with the click of a button. Power system component data is stored by using the integrated graphical capabilities of the power system analysis software. This information can be obtained in various ways for instance, equipment nameplate, manufacturer catalogues, drawings, calculations and actual measurements.

With the aid of the PSAS a complete model of the plant can be built. The software uses standard power system models to model each component in the system.

Once the complete system is modeled and all system parameters are stored, the Power system database can be integrated to the rest of the system as described in the Figure 1.

An object orientated ASCII file or DOLE interface (Direct Object Linking and Embedding) can be updated with the latest real time system status. This information is obtained from the data acquisition system by means of instantaneous plant measurements. The DOLE interface is a common ASCII interface for all kind of conventions. Creating new objects, changing values of existing objects and drawing single line graphics are the main features of the interface.

The PSAS can be operated in an on – line mode with a direct connection to a SCADA system. This feature enables to access real measured system data, either the last actual ones, or any set, which has been saved previously by the SQL server

By activating the last measured system status, which will comprise generation, loads, transformers tap settings,

shunt reactor switching and system configuration, this option allows the analysis of actual system operation with respect to:

- Active and reactive power flow, voltage profile, system losses and reactive power compensation requirements
- General fault analysis:
- System losses optimization;
- Spinning reserve and load shedding supervision and optimization;
- Analysis of measures for system oscillation stabilization.

The access to previously stored data allows the analysis of any occurred fault or otherwise specific situation needed to be studied. In this case, the main advantage of accessing SCADA history files is the ability of updating power system data without being forced to edit any data manually.

The procedure implemented for SCADA interfacing is as follows:

1. The operated power system must be set up in the PSAS using the single line graphic editor or the data manager. This configuration must reflect the general structure of the operated system, including the existing busbars (substations) installed lines, cables, transformers, shunt-reactors etc. Specific operation information like loads, generation or any general "on/off" information concerning bus-couplers, circuit breakers or isolators will later be overwritten.
2. Data transfer files have to be generated, organized and stored periodically by a SCADA interfacing program, which may run on the PSAS computer system or the SCADA system itself. The generated update files have to follow a certain structure and file naming convention. In general, they can be best described as "incremental files" containing the operated system actual data.

The results of the studies can be verified in different ways depending on the type of study. The user of this system should have a good knowledge of power system calculations and should be able to perform most of the calculations manually.

The user can update the PSAS with the latest system data or historical data depending on the type of study to be performed.

3. System studies

Power system studies properly conceived and conducted, are a cost – effective way to prevent surprises and to

optimize equipment selection. In the design stage, the studies identify and avoid potential deficiencies in the system before it goes into operation.

In the existing systems, the studies locate the cause of equipment failure and misoperation and determine corrective measures for improving system performance.

The complexity of modern industrial power system make studies difficult, tedious, and time consuming to perform manually.

The following section address the most common studies that can be performed with the PSAS for the design and operation of an industrial or commercial power system [3]:

- **Load flow studies.** Load flow studies determines the voltage, current, active, and reactive power and power factor in a power system. Load flow studies are an excellent tool for system planning.
- **Short Circuit Analysis.** Short - circuit studies are done to determine the magnitude of the currents flowing throughout the power system at various time intervals after a fault occurs. The magnitude of the currents flowing through the power system after a fault varies with time until they reach a steady – state condition. The information in short – circuit analysis is used to select fuses, breakers, and switchgear ratings in addition to setting protective relays.
- **Stability studies.** The ability power system, containing two or more synchronous machines, to continue to operate after a change occurs on the system is a measure of stability. The stability problem takes two forms: steady state and transient state. These types of studies are probably the single most complex ones done in a power system. A simulation will include synchronous generator models with their controls, i.e., voltage regulators, excitation systems, and governors.
- **Harmonic Analysis.** Harmonic Analysis is commonly done to predict distortion levels for addition of a new harmonic – producing load and capacitor bank. Only very small circuits can be effectively analyzed without a computer program. Typically the PSAS for harmonic analysis will provide the engineer with the capability to predict the actual distortion based on models of converters, arc furnaces, and other nonlinear loads.
- **Protective Device Coordination.** The objective of a protection scheme in a power system is to minimize hazards to personnel and equipment while causing the least disruption of power service. With the advent personal computer, the light – table

approach to protective device coordination is being replaced by PSAS with visual representation of device coordination [4].

4. Requirements and qualities of good power system analysis software

In an industrial plant where continuity of supply is of vital importance for the production process or in city municipalities and electricity suppliers where continuing system planning and feasibility studies is performed this system is a solution to quick, effective and accurate system studies.

The PSAS used should be an advanced, fully interactive simulator package, dedicated to electrical power system analysis in order to achieve the objectives of planning and operation optimization.

It also should be both, a large - scale power system analysis package as well as a "pocket calculator" for solving electrical engineering routine problems.

Nowadays, there exist many commercial packages capable of addressing most of the power system analysis problems. However, these packages differ considerably in the quality of the results and in their computational efficiency due to various approaches adopted to tackle the following essential items:

- General design of the package
- Data structures and data base implementations
- Problem definition
- Mathematical formulation
- Solution techniques
- Programming methodology
- Portability and maintainability
- Man-machine communication structures and techniques
- Versatility of the package
- Package interactions
- Comprehensive multi-level network models on a physical, device-based description; provision of general user interface for model definition.
- Enhanced tabular and graphic output - already designed for reports and documentation purposes.
- Utilities for parameter calculations (OHLs, cables, machines, etc.) "
- Comprehensive dynamics and relay model library with provision for user-defined macros and models.
- Windows-operation, Interactive, batch mixed mode operation or with historical plant data that is stored in the SCADA data tables and relational database.

5. Conclusion

The planning, design, and operation of industrial and/or commercial power systems require several studies to

assist in the evaluation of the initial and future system performance, system reliability, safety, and the ability to grow with production and/or operating requirement studies.

Integrating PSAS in the information chain used in industry has obvious advantages for system engineers in terms of system studies, planning and training of personnel.

The advantages are:

- Access to system component constants with the push of a button.
- Real time and Historical system studies.
- Network planning for future extensions
- Training of key personnel.
- Cost saving - Advance system studies can be done in-house. It is therefore not necessary to pay a lot of money to consultants.

Some Drawbacks:

- To install and to make full use of the system the information structure as described has to be in place before PSAS can be integrated. This process can be very costly for companies who do not have the complete control hierarchical structure in place.

5. References

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HOW TO CHECK THE RELIABILITY OF A MODEL IN POWER SYSTEMS

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ABSTRACT

The Theory of Dynamical Systems (TDS) indicates that Structural Stability (SS) plays an important role in explaining the behaviour of a non-linear system. As the Power System is in our days highly a non-linear system, it is appropriate to explain the role of SS in this field. A short introduction of the concept will be given in the context of Power System. A single machine infinite bus system will be analysed with different kinds of perturbations (linear, non-linear and time perturbation). The full phase space of the system is plotted and SS evaluated. Then, a load is introduced in order to study the relationship between SS and load modelling. This example demonstrates the fact that the load influences the SS of a system. Finally, will be shown that load modelling is crucial for a proper analysis of SS. An intimate relationship among SS, bifurcation and voltage collapse justifies the use of SS.

1. INTRODUCTION

Economical driving forces are pushing the utilities to transfer larger quantities of power to meet the demands. Due to the high construction costs of power transmission lines and of power generating stations, power companies are delaying the construction and, instead, are trying to develop new methods for increasing and improving the possibilities of the actual network. It is believed that deeper understanding of the problems that are underpinning the stability of the system could contribute to these goals.

In this paper, a new kind of stability, structural stability [1] [2], is studied. Its significance in the context of load modelling is examined. It is shown

that beyond certain values of load parameters, in the case of static models, the system may become structurally instable. The immediate application of these results is in the voltage stability problems [3] [4].

Broadly speaking, structural stability refers to a domain in the parameter space such that for small variations in the parameter vector inside the domain, the phase portrait does not change qualitatively. In Lyapunov stability, we similarly talk about the region of attraction of an equilibrium point in the phase space such that for any initial condition in the region, the trajectory approaches the equilibrium point asymptotically. While considerable research has been done in the Lyapunov stability area, comparatively very little has been done in the structural stability area.

2. STRUCTURAL STABILITY

2.1 Mathematical Introduction

The mathematical description of any real system or physical phenomenon is always imprecise because of inevitable errors in determination of the parameter values or because of neglect of minor factors. The mathematical description (model) has no practical value unless it is insensitive to small changes or perturbations, or otherwise there is no guarantee that the conclusions based on the model are consistent with the behaviour and properties of the original system. This insensitivity is referred to as structural stability (SS).

Power System dynamic analysis is based on non-linear differential equations (as the swing equation for a classical machine model). Other components of the system are described by algebraic equations (for example the network

equations). In general, we are dealing with a differential-algebraic system of equations, which for the sake of simplicity we will denote:

$$\dot{x} = f(x, u) \quad (2.1)$$

$$v = (x, u) \quad (2.2)$$

where x is the state variable vector, u is the input vector and v is the output vector. The set of equations (2.1) and (2.2) is the basic starting point in any study performed on a system.

The concept of SS for a phase system was introduced by Andropov and Pontryagin (1937) [5]. Two complete characterisations for a planar system have been given: by Debaggis [6] and the other by Peixoto [7].

2.2 Debaggis' definition

Let G be the region where $f(x, u)$ has continuous first partial derivatives and let $M(x_1, u_1)$ denote a point in this region. Consider the perturbed system:

$$\dot{x} = f(x, u) + p(x, u) \quad (2.3)$$

where $p(x, u)$ belongs to C^1 for all points $M(x_1, u_1)$ in G .

Definition: let $\rho(M_1, M_2)$ denote the Euclidean distance between two points in the space. The system (2.1)-(2.2) is said to be SS in the region G if for each $\varepsilon > 0$ there exists a $\delta(\varepsilon) > 0$ such that for all $p(x, u)$ which belong to a class C^1 and satisfy the condition

$$|p_i| < \delta \quad \frac{\partial p_i}{\partial x_j} < \delta \quad (i = 1, n \quad j = 1, n) \quad (2.4)$$

there exists a topological transformation T of G onto itself with the additional properties:

- (a) $\rho[M, T(M)] < \varepsilon$
- (b) T maps trajectories of (2.1) onto trajectories of (2.3)

The following properties characterise a structurally stable system in two dimensions:

- (a) the critical (equilibrium) points of a structurally stable system (2.1) can only be nodes, foci and saddle points.
- (b) a separatrix of a structurally stable system (2.1) cannot start from (end in) a saddle point and terminate on (start from) another saddle point in G (heteroclinic orbit).
- (c) in a structurally stable system, a separatrix that starts from a saddle point cannot converge toward the same saddle point (homoclinic orbit).

- (d) a structurally stable system (2.1) has only a finite number of limit cycles in G .

There are some more properties, which describe the SS in, 2-d but they have little importance for the scope of the present paper.

2.3 Peixoto's theorem

Using the system of equations (2.1 – 2.3), let f be a C^1 vector field on a compact, 2-d, differentiable manifold M . Then, f is SS on M if and only if:

- 1) The number of critical points and periodic orbits is finite and each point is hyperbolic.
- 2) There are no trajectories connecting saddle points.
- 3) The nonwandering set Ω consists only of critical points and limit cycles.

These two definitions of SS in planar system imply that a structurally stable system will contain, in its own field f , only sinks, saddles and sources points, and repelling and attracting orbits.

After Peixoto's theorem, it was thought that it can be extended to N -dimensional systems ($N \geq 3$) but later studies indicated that this was not true. When systems with $N \geq 3$ are considered, we do not have an exact criterion to evaluate the SS of that system.

Because of this lack of criteria for higher dimension system, we will analyse in the next two sections, two base cases which have the lowest phase space dimension. Possible approaches for real power system (real having here the meaning of hundreds of buses) will be given in conclusion.

3. SIMPLE MACHINE INFINITE BUS SYSTEM (SMIB)

The SMIB (figure1) system is described by the well-known dynamic swing equation. In its simplified form, it reads:

$$M \cdot \frac{d^2 \delta_m}{dt^2} + D \cdot \frac{d\delta}{dt} = P_m - P_e \quad (3.1)$$

where M is the angular momentum of the rotor, D is the modified damping coefficient, δ_m is the rotor angle in radians with respect to a stationary axis, P_m is the mechanical power and P_e is the output electrical power, $P_e = \text{Re}(E\bar{I})$, where E is the busbar voltage and I is the current.

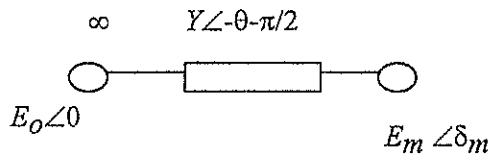


Fig.1

From figure 1 one can see that:

$$P_e = -E_o E_m Y \sin(0 - \delta_m + \theta) \quad (3.2)$$

where Y and θ are the magnitude and the phase of the admittance between the generator and the infinite bus, E_o and 0 are the magnitude and the phase of the infinite bus and E_m and δ_m are the magnitude and phase of the generator. After the substitution of (3.2) into (3.1) we get:

$$M \ddot{\delta} + D \dot{\delta} = P_m - E_o E_m Y \sin(\delta - \theta) \quad (3.3)$$

The non-linearity of the system is due to the presence of the *sine* function.

We will apply three different perturbations to the system – a linear, a non-linear and a time dependent perturbation. The phase space of the

system, with different kinds of perturbations is represented in figures 2a-d.

The system of differential equations used to construct the phase space is

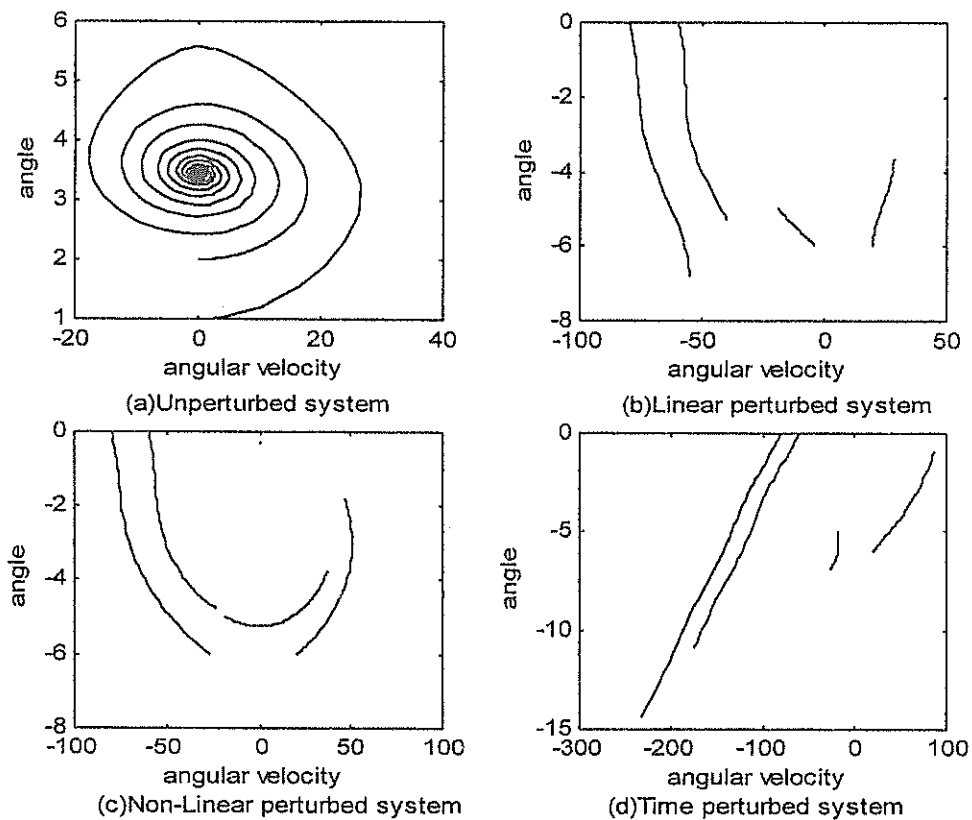
$$\frac{d\delta_m}{dt} = \omega \quad (3.4)$$

$$M \frac{d\omega}{dt} = -D\omega + P_m - YE_o E_m \sin(\delta - \theta) \quad (3.5)$$

One can see that the angle is oscillating around a fixed value ($\delta \approx 3.5$) as the angular speed oscillates. The system is stable at this stage. Two trajectories have been represented in the phase space. They have different initial conditions but as the time passes, they come together, moving towards an equilibrium point.

Let us now apply a linear perturbation on the system. We will add a factor $\varepsilon \delta$ to the equation (3.5), where $\varepsilon = 0.2$. The new phase space is plotted in figure 2.b. Two nodes appear and all the trajectories are ending up in them. This indicates that the points are stability points. In terms of

Fig.2 The values of the parameters used are: $M = 0.0138$; $D = 0.5$; $P_m = 0.91$; $E_o \cdot E_m = 3.02$; $\theta = 0$.



Peixoto's theorem, this phase portrait is not topological equivalent with the one from figure 2.a because new critical points were created.

Applying another type of perturbation, a non-linear one- $\varepsilon\delta^2$, the phase portrait is shown in figure 2.c. The four trajectories which are plotted here respect no pattern. Clearly the phase portrait has changed again. This suggests that our system is not structurally stable.

Finally, let us perturb the system using a time dependent term $\varepsilon \cdot \omega \cdot \cos t$. The phase space is constructed again (figure 2.d) and as in the previous case, the trajectory does not respect any pattern. In this case we can say that the figures 2.c and 2.d are topologically equivalent but they correspond to a perturbed system and they are not topological equivalent to the unperturbed system, as SS requires.

We end this section with a few remarks:

- The system analysed here is not SS because the unperturbed system's phase space cannot be mapped in any perturbed system phase space.
- The word "perturbation" used in this paper, has the meaning of a mathematical perturbation, i.e an equation acquire a new term which depends on the one of the state variables.
- This does not mean that the SMIB system is inappropriate for studies but rather that the perturbations used (linear, non-linear and time dependant) were very severe and unrealistic. The swing equation does not admit the terms that were added. At a closer inspection, no term is allowed to be added to the swing equation.
- The reason behind our choice of unrealistic perturbations was to illustrate what one should expect from an unstable system.
- A natural and appropriate way to perturb the structure of a power system should be found.

4. A SMALL POWER SYSTEM

In this second example we will use the changed version of the previous machine. We keep the infinite bus and generator and we add a load in between. The idea is to introduce the active and reactive powers and to be able to study the

influence of the structure of the load on SS. The power system is represented in figure 3.

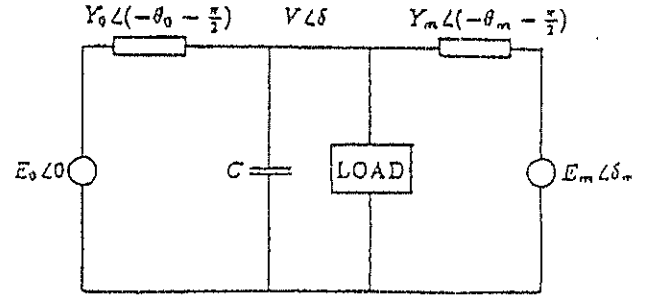


Fig. 3

We assume that the voltage E_m at the generator is constant so only V and δ are allowed to change in time (V and δ are the voltage and the phase at the load bus). The system of equations which describes this power system is now:

$$\dot{\delta} = f(\delta, V, \delta_m, \omega) \quad (4.1)$$

$$\dot{V} = g(\delta, V, \delta_m, \omega) \quad (4.2)$$

$$\dot{\delta}_m = h(\delta, V, \delta_m, \omega) \quad (4.3)$$

$$\dot{\omega} = k(\delta, V, \delta_m, \omega) \quad (4.4)$$

We have kept the equations (3.4) and (3.5) and we add the equations that describe the loads:

$$P_s = (p_1 + p_2 V + p_3 V^2) P_1$$

$$Q_s = (q_1 + q_2 V + q_3 V^2) Q_1$$

where P_1 and Q_1 are the active and reactive of the PQ loads, p_i and q_i are constants and the subscript 's' stands for static. This model is commonly referred to as the ZIP model because it is composed of a constant impedance (Z), a constant current (I) and a constant power (P) component.

Using the ZIP model, by changing the constants p_i and q_i , for example from any value to 0, the structure of the power system is affected. By "structure" we understand the relationship among the state variables of the system of equations. In this manner we are changing the structure in a natural way, as in real life the structure of the power is changing, and we avoid running into artificial problems as it happened in the previous example, when the system was perturbed arbitrarily.

In this case the phase space is a 4-d. We cannot visualise it as a whole, so we chose a projection in the $(V - \omega)$ plane. The phase spaces for different situations are plotted in figure 4a-f.

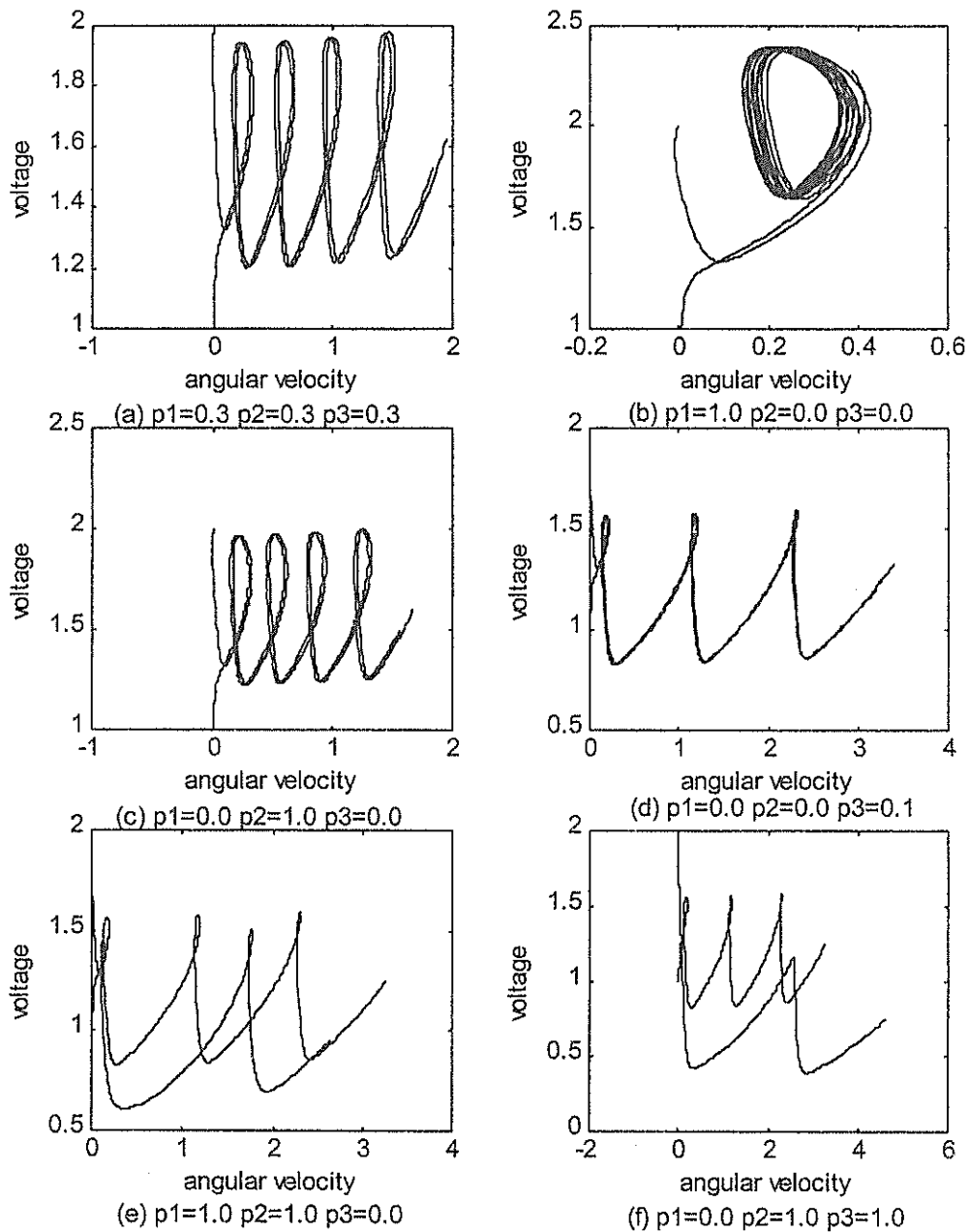


Fig. 4; The phase space of the small power system

We start with the parameters of load $p_i = q_i = 0.3$, a mixture of constant P , constant I and constant Z . Two trajectories are calculated with different initial conditions. The phase space is shown in figure 4.a. One can see that even if the trajectories start in opposite zones of the phase plane, they come together very quickly and the voltage oscillates around a constant value of

1.6. The parameters have been chosen so that the voltage is oscillating a little, in order to see a change in the phase space.

Next, the mixture of constant P , constant I and constant Z is changed as follows:

$p_1 = q_1 = 1.0, p_2 = q_2 = 0.0, p_3 = q_3 = 0.0$;
as in figure 4.b

$p_1 = q_1 = 0.0, p_2 = q_2 = 1.0, p_3 = q_3 = 0.0$;
as in figure4.c

$p_1 = q_1 = 0.0, p_2 = q_2 = 0.0, p_3 = q_3 = 1.0$;
as in figure4.d

$p_1 = q_1 = 1.0, p_2 = q_2 = 1.0, p_3 = q_3 = 0.0$;
as in figure4.e

$p_1 = q_1 = 0.0, p_2 = q_2 = 1.0, p_3 = q_3 = 1.0$;
as in figure4.f

Looking at all the phase portraits, we conclude that they are topological equivalent (by simple inspection), so that the system is SS from the load point of view.

By changing the parameters (p_i, q_i) , we changed the structure of the power system, (structure in the mathematical sense) as required by SS. The response of the system is the same under different load conditions.

In summary, for the particular values of the parameters used, the system exhibit a slight instability, voltage is not constant but oscillates around an equilibrium point. Instead, the system is at SS even if the structure of the load is modified in a substantially manner from a mixture of constant P , constant I and constant Z (see figure 4.a) to a pure constant power load (see figure 4.b). The results allow us to conclude that the model used is appropriate for doing studies on it.

5. CONCLUSION

For each case a conclusion was drawn. How the SS can be evaluated and how can it be applied to a small power system was discussed.

For a real power system (hundreds or thousands of buses) the problem of representing the phase space becomes impossible. The dimension of the phase space is huge and no visualisation is possible. Still, there is a way to visualise part of the phase space by choosing a projection in plane. The state variables necessary for the plane projection are identified using the participation factor.

The authors want to make only one final comment about SS. There is some confusion in the literature when the problem of SS is studied in conjunction with the load modelling. One could object that changing the load parameters (p_i, q_i) in a power system is something natural, happening

any time. Why do we then associate this change with SS? In our view, we deal with two different problems.

1. One can study a system to see how it responds to changes in load modelling. At initial time t_0 , one set of parameters is used - it is referred to as LM1. Later, at t_1 , the distribution is changed to LM2 and so on. This classical procedure employed by a power system has nothing to do with SS.
2. What one has to do in order to study SS is the following: the same system as above could be studied with the load LM1 for a period of time. If the system is stable, the study is stopped. Then a new load LM2 is used, and the study of the system is done at t_0 , using the same initial conditions as before. The procedure could be repeated for different loads LM. When SS is studied, the same initial conditions at the same initial time are used.

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MICRO-CONTROLLER BASED DYNAMOMETER

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Abstract:

The dynamometer is a device used to measure the torque of a machine, whether that machine is a diesel engine or an electrical motor. In the Faculty of Electrical Engineering, P.E. Technikon, only motors are dealt with. Since most dynamometer units are imported, and the exchange rates are poor, these units are becoming further out of reach of educational institutions. Even more so with decreasing budgets. Therefore the need has arisen to develop an affordable dynamometer that is to be used in the laboratories of educational institutions.

Keywords:

Micro-controller
Strain-gauges
DC-injection brake

1. INTRODUCTION

Torque is a concept that is not readily understood by the average student. The theory behind torque is learnt in the classroom, and how the torque-speed characteristics of different motors vary. Displaying this practically requires specialised laboratory equipment, which will cement the students' understanding of the torque-speed characteristics of motors.

Due to increased student turnover, more laboratory equipment is needed to successfully run practicals.

At Port Elizabeth Technikon, the budget was no longer sufficiently large to purchase the dynamometers needed to maintain the high standard of practical training in the Power Systems Department. This has made it necessary for a cost-effective dynamometer to be developed.

1.1 Sub-problems

1.1.1 Available dynamometers are extremely costly.

1.1.2 Most of the available dynamometers don't have all the functions which are preferred, such as temperature monitoring of the motor being tested.

1.2 Solutions

It has been concluded that the solution is to develop a dynamometer that is designed around the Intel 80C196KC micro-controller, for the following reasons:

1.2.1 The cost of a micro-controller based dynamometer is significantly less than the other available units.

1.2.2 Some of the available units don't provide an indicator for speed. This can be overcome by using a micro-controller based dynamometer, where many more transducers can be linked together, giving a much greater flexibility.

1.2.3 The micro-controller based dynamometer will be much easier to use, since all the readings will be displayed on the liquid-crystal display simultaneously

1.3 Objectives

1.3.1 By means of transducers, the following values of the driving motor are to be measured, and once processed are to be displayed on the LCD:

- 1.3.1.1. current.
- 1.3.1.2 temperature.
- 1.3.1.3 speed.
- 1.3.1.4 torque.

1.3.2 The brake must be capable of stopping a motor with a power rating of up to 1 kW.

1.3.3 The dynamometer must be capable of measuring up to 7.5 Nm of torque.

1.3.4 The dynamometer must be capable of measuring up to 10 A of motor current.

1.3.6 The brake is a 3-phase motor with the following characteristics:

- 1.3.6.1 power rating: 2.2 kW.
- 1.3.6.2 number of poles: 6.
- 1.3.6.3 type: squirrel-cage induction.

2. DESIGN AND DEVELOPMENT

2.1 Choice of transducers

2.1.1 A Hall-effect current transducer was chosen because it has a voltage output which is proportional to the current; and the analog-to-digital converter works on voltage inputs. The Hall-effect current transducer is also compact, compared with a conventional current transformer.

2.1.2 A temperature transducer was chosen to be affixed directly to the motor's windings to provide an accurate measurement of the winding temperature. (Reference 9.)

The transducer that was chosen, was done so for the following reasons:

2.1.2.1 it is very small, which is ideal for this application, seeing as there is limited space between the windings and the motor housing.

2.1.2.2 it is relatively inexpensive for a temperature-measuring device and is significantly cheaper than similar devices made by other manufacturers.

2.1.2.3 it requires no calibration and provides a fixed voltage output that is proportional to the temperature.

2.1.3 for the speed-measurement of the driving motor, an optical coupler is used in conjunction with a disc that has holes around the circumference.

The optical coupler that was chosen was done so for the following reasons:

2.1.3.1 it has no direct contact with the motor or the coupling, thus being maintenance-free.

2.1.3.2 it has a Schmitt-trigger output which is well matched to the high-speed digital input on the 80C196KC micro-controller. (Reference 6.)

2.1.4 To measure the torque of the motor, strain-gauges have been bonded to a specially-constructed bracket, onto which the brake is mounted. The strain-gauges have been connected to form a Wheatstone bridge circuit, that provides a voltage output that corresponds to the torque of the driving motor.

Strain-gauges have been chosen instead of load-cells because load-cells which would be suitable for this application are far too expensive.

2.2 Choice of brake design

Through laboratory experimentation, it was found that if the DC-injection principal could be applied to a 3-phase motor, it could operate quite efficiently as a brake. However, the brake can draw up to 10 A of current to bring a 0.75 kW motor to standstill. This means that a fairly powerful means of control is needed. The following are some methods of control that were investigated:

2.2.1 simple phase-control using a TRIAC circuit. (Reference 1, 2, 7.)

2.2.2 use of an auto-transformer.

2.2.3 PWM control using a power-semiconductor as the switch.

2.2.4 although the phase-control method is suitable for resistive loads, it is very unstable with an inductive load like the brake. This is due to the trigger circuit taking the voltage as the trigger reference, while in actual fact TRIAC-triggering needs to be based on the current. (Reference 7.)

The PWM control method was opted for because of the following pitfalls of the other 2 methods:

2.2.5 auto-transformers, or variacs, are very expensive and bulky, for the amount of current needed by the brake.

2.2.6 the PWM control unit is very cost-effective and provides a stable control over the amount of DC current injected into the brake. It is also a very compact arrangement. (Reference 9.)

2.3 Choice of micro-controller

The 80C196KC micro-controller was successfully used in the single-phase phasor meter that was developed last year at Port Elizabeth Technikon. (Reference 10.)

The 80C196KC is a 16-bit micro-controller, with a 12-bit on-board analog-to-digital converter. (Reference 6.) It was decided that this analog-to-digital converter would be sufficient for the project.

2.4 Overview of software

The analog signals from the following transducers are fed into the analog-to-digital converter after being amplified and signal-conditioned:

2.4.1 current

2.4.2 temperature

2.4.3 torque

This data is then manipulated by software so that it can be displayed in a useful manner.

The digital signal from the speed transducer is fed in via the High Speed Input module on the 80C196KC micro-controller. This digital signal is used to clock a timer, and a time-reference is created by the use of interrupts. (Reference 6.) From the time reference, the speed of the motor can be calculated.

3. CONCLUSIONS

The micro-controller based dynamometer will meet the needs of educational institutions, because it is cost-effective and will be perfect for demonstration purposes in the electrical laboratories of P.E. Technikon.

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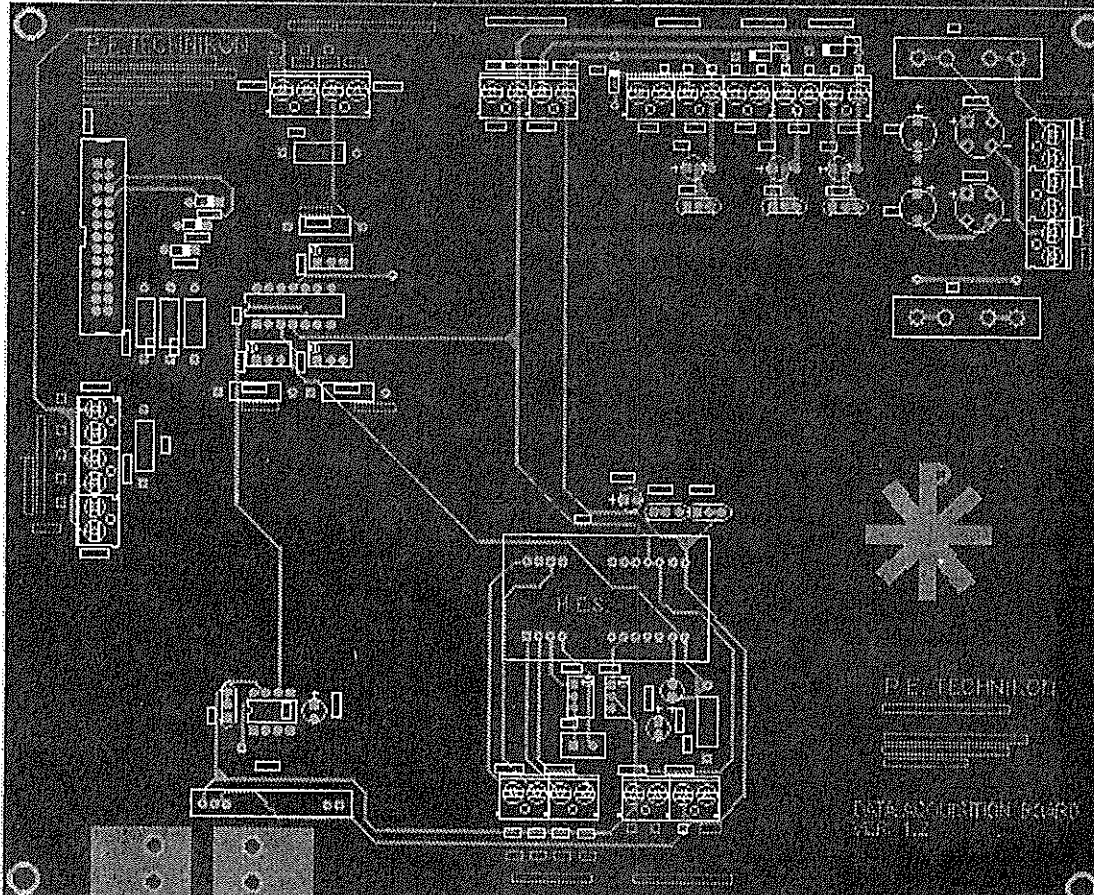
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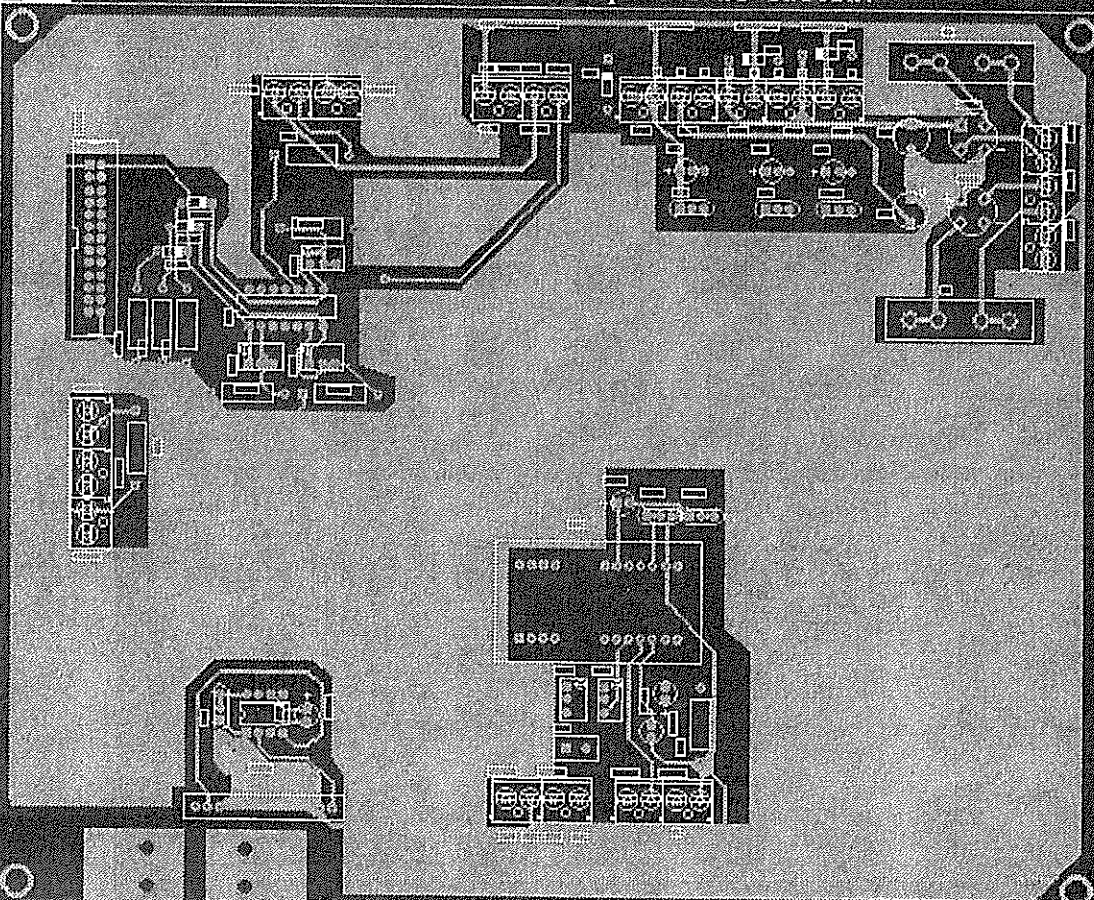


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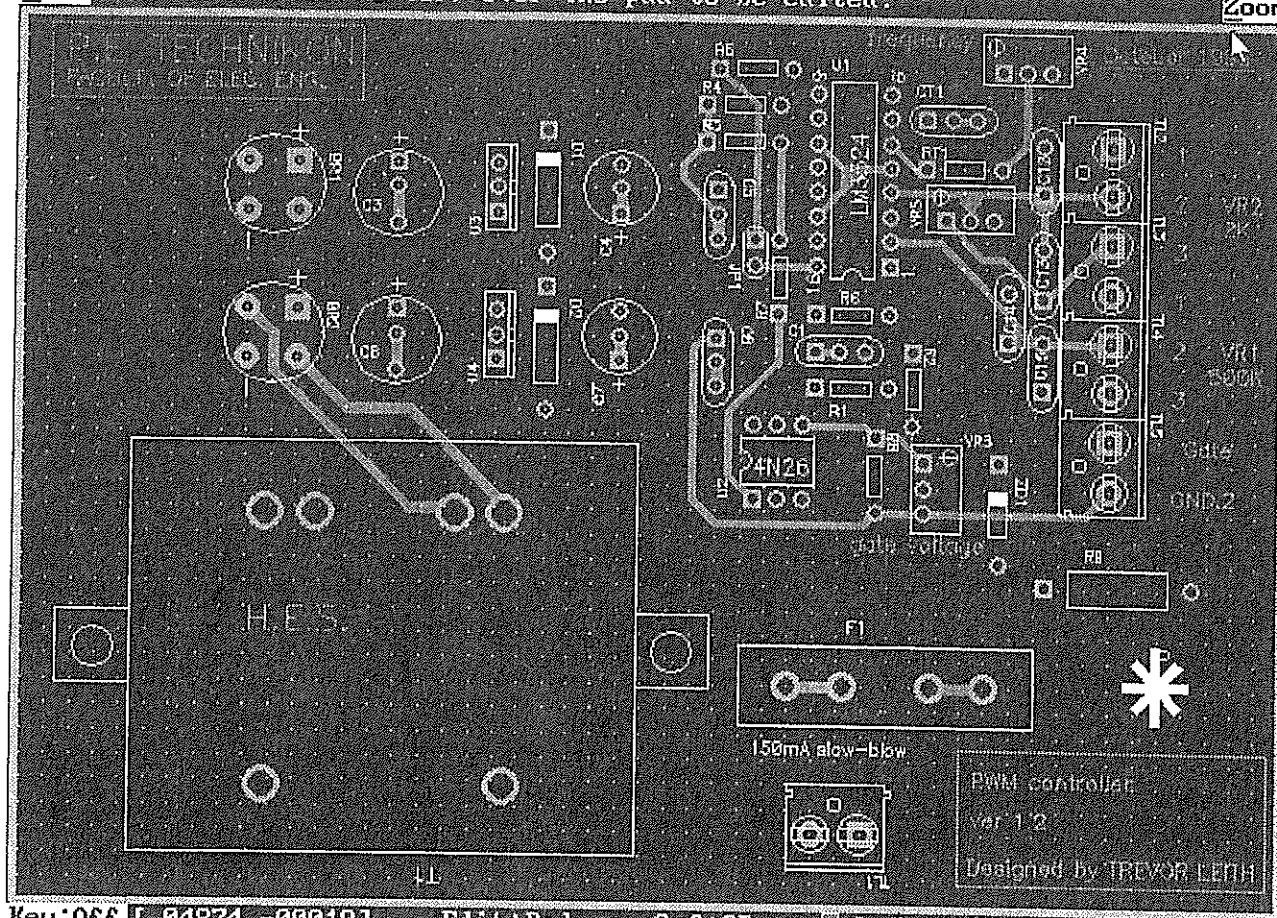


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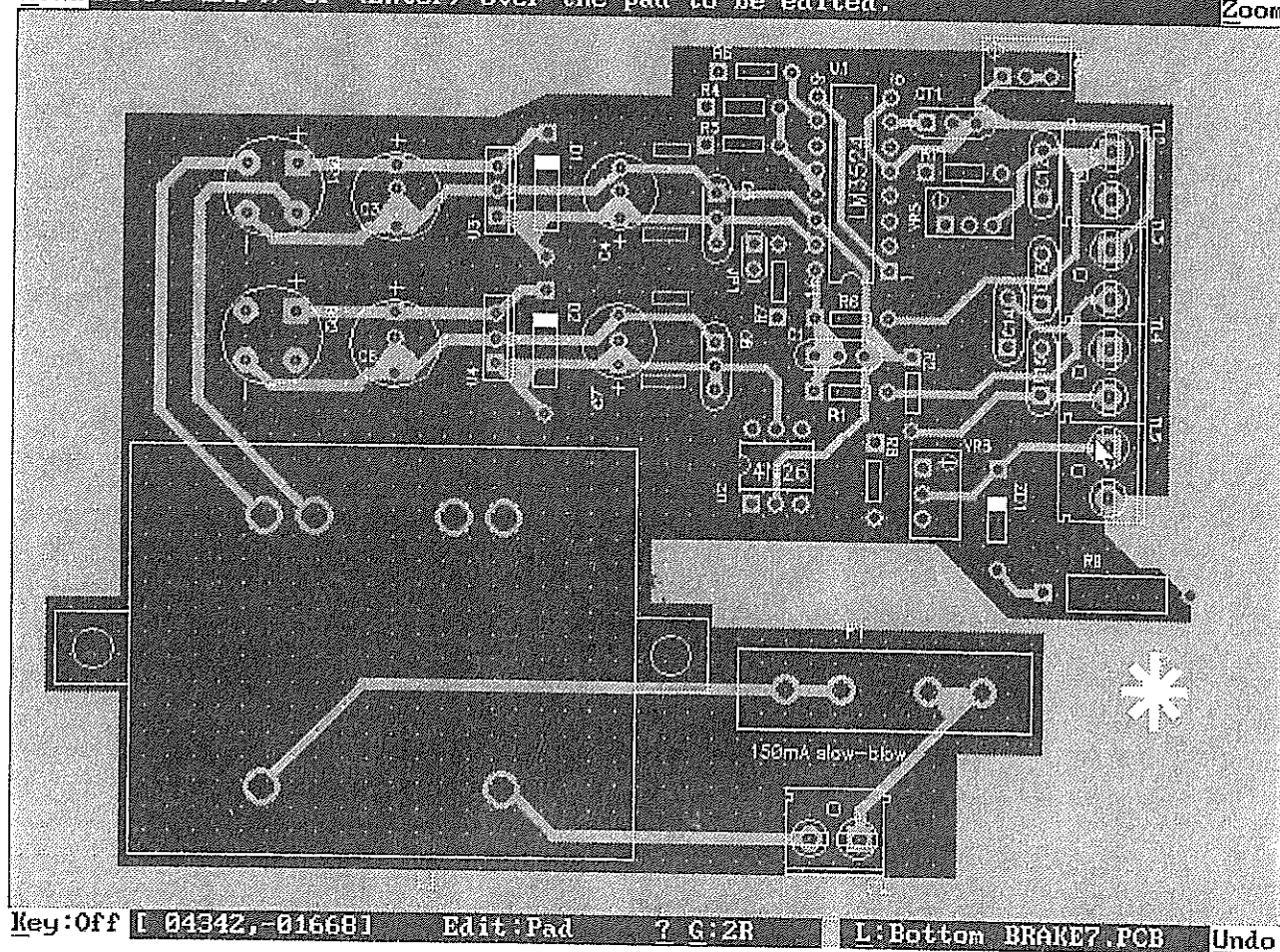
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A COST COMPARISON BETWEEN THE INSTALLATION OF STORAGE WATER HEATERS AND INSTANTANEOUS WATER HEATERS ON A THREE-PHASE LOW VOLTAGE FEEDER

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Abstract: There have been suggestions for the wide spread use of instantaneous water heaters(IWHs) for domestic use in South Africa. This paper describes an examination of the effects that IWHs will have on the reticulation system. In order to examine the voltage conditions on a reticulation system an Energy Block Model was developed. The Energy Block Model was used to derive a planning level of load currents, using the Binomial Random Theorem and to compare losses. Voltage drop and losses on a typical feeder were calculated using the Herman Beta algorithm and by simulation. The extra costs of reticulation needed to supply IWHs exceeds the savings achieved by the user, so that the overall cost of IWHs is higher than that of conventional storage water heaters.

Keywords: instantaneous water heaters, reticulation, energy block model

1 INTRODUCTION

Water heating accounts for 40% to 60% of the South African domestic electrical usage [1]. Any method of making water heating more efficient would therefore play a significant role in reducing South Africa's energy consumption. The cost of building new power plants and the running costs of peaking plants will then decrease, reducing the cost of electricity for all South Africans.

Instantaneous water heaters effectively reduce standby losses. This makes them 14% more efficient than storage water heaters [2]. This improved efficiency is the reason why many manufacturers and suppliers of instantaneous water heaters are trying to get three-phase service connections installed in order to cater for these heaters.

This paper describes an examination of the effects of storage heaters and instantaneous water heaters on the load profile of a group of customers and what feeder resistance is needed to ensure that the voltage drop of both heating systems is comparable.

1.1 Alternative heaters

For the purposes of this study it was important to ensure that the instantaneous water heaters and storage water heaters or hot water cylinders (HWCs) used had similar hot water delivery performance.

This meant that for every hot water event the same flow rate and temperature had to be delivered by both the storage and instantaneous water heaters. In order for this to be true the same amount of energy had to be used by both water heaters. The water delivery performance of a 21kW instantaneous heater was

similar to a storage water heater with a 3kW element and was therefore used for this study.

1.2 Consumer types

According to the NRS-034 standards, a consumer with electrical water heating is classified to be in the high or very high consumption class [3]. For the purposes of this study, the consumers from Die Bos, a suburb in Helderberg, were taken as representative of these consumer classes.

Their load profiles were the basis for calculating the voltage drop in feeders, using the Herman Beta algorithm [4] and a computer simulation.

1.3 Benchmark feeders used

Benchmark feeders based on those described by Gaunt and Sellick were used to compare the two heating systems[5]. These represent consumer densities of 16 and 8 consumers/ha.

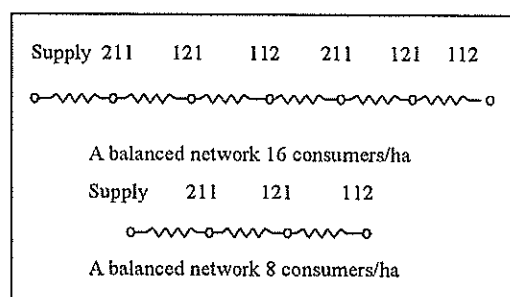


Figure 1: Benchmark feeders used

2 ENERGY BLOCK MODEL

2.1 Equal energy concept

For the water use incident (water event), the energy required by each type of water heater is the same. This can be represented as an energy block. However, one energy block uses low power over an extended period and the other energy block uses greater power over a shorter duration.

As outlined in section 1.1, the instantaneous water heaters has a power rating of 21kW while the storage water heater has a power rating of 3kW. Since 21kW is seven times 3kW, the 'on' duration of a instantaneous heater is one seventh of that of the storage heaters, for the energy block to be of equal magnitude.

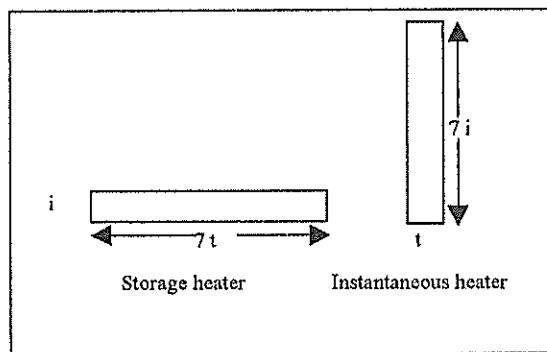


Figure 2: Energy Block Model

2.2 Binomial Random Theorem applied to Energy Block Model

The Binomial Random Theorem is used when a random variable is either in the 'on' or 'off' state. The Binomial Random Theorem was therefore used to estimate how many heaters, out of a total n , were on at the same time. A normal distribution was obtained from which the 97% confidence level was used to derive the highest current expected for the two heating systems, termed the design current used for feeder planning.

Figures 3 and 4 illustrate the average current for groups of consumers (97% confidence) and the associated coincidence factors.

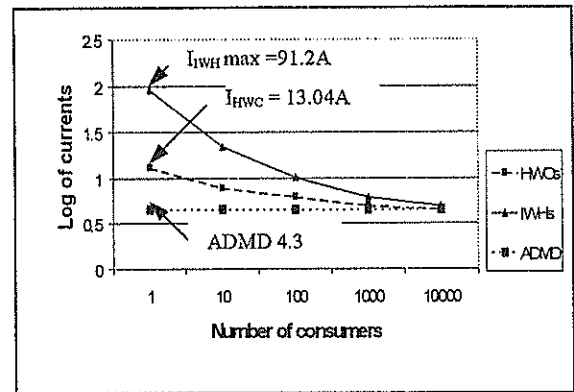


Figure 3: Average current per consumer

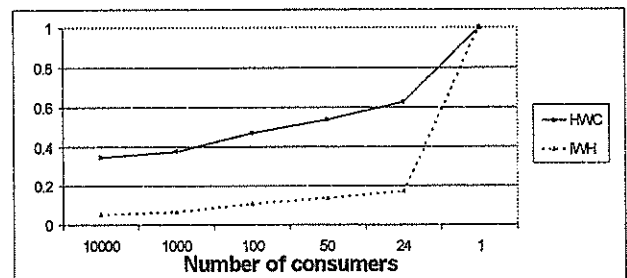


Figure 4: Coincidence factor for heaters

These graphs indicate that although the coincidence factor was lower for the instantaneous water heaters, the design current was still higher for the IWHs. This finding contradicted previously published statements that the same design current can be used for comparing the different water heating methods [7].

3 FEEDER RESISTANCE

Two methods were used to calculate the feeder resistance. A lower resistance means a thicker cable and therefore a more costly cable and affects the losses in the network.

3.1 Herman Beta Method for calculating the voltage drop

The measured currents taken from Die Bos, Helderberg were used to represent typical consumers with HWCs. The beta parameters were calculated at the time of the system maximum demand.

The load profiles of each consumer were examined for the period close to the system peak and water events identified. These were replaced with equivalent energy use for an IWH, and a new load profile constructed for the period. The transformed load currents at the time of the system peak provided a distribution of currents for a group of consumers equivalent to a probability density function (pdf), for which beta parameters were calculated.

Table 1: Equivalent parameters IWH/HWC based on the current reading of HWC consumers at Die Bos

	α	β	μ	σ	c_b
HWC	2.98	3.19	18.4	7.12	38.2
IWH	0.3272	0.6813	12.97	13.2	38.16

The Herman Beta Method was then used to find the resistance of two benchmark feeders needed to ensure that with both group profiles the voltage drop was the same.

The feeder resistances from the Herman Beta Method are shown below.

Table 2: Resistance values from Herman Beta Method.

Densities:	Resistance Ω/km	
	HWC	IWH
8 consumers/ha	0.59	0.39
16 consumers/hs	0.42	0.28

These results indicate that the reticulation system for the current profile of the instantaneous water heaters needs significantly larger cable sizes, thereby making the system more expensive.

3.2 Computer simulation of voltage drop on feeder

The distribution of consumer loads were not visually similar to conventional beta pdfs, so a computer simulation was used to check the voltage drops determined using the Herman Beta algorithm.

The simulation randomly selected and placed 24 loads from 61 consumers at the nodes of the benchmark feeders.

The results from this method closely matched those of the Herman Beta Method; the results are presented in the table below. This indicates that the data fitted the Beta Distribution

Table 3: Results from the computer simulation

Densities:	Resistance Ω/km	
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8 consumers/ha	0.58	0.39
16 consumers/hs	0.432	0.28

3.3 Losses based on Energy Block Model

The formula i^2Rt was used to calculate the losses using the values for i and t from the Energy Block Model. The resistance of the feeder and service connections were calculated, and were not the same for the two types of water heating.

The losses of a three-phase feeder per water event are therefore:

For instantaneous heaters:

$$= 3 \times (7/3 \times i^2) \times R_i \times \text{length} \times t$$

For storage heaters:

$$= 3 \times (1/3 \times i^2) \times R_s \times \text{length} \times 7t$$

The losses, because the length of the feeders is the same, are almost two times greater on the instantaneous heating system.

A similar comparison was done for the service connections and again the losses on the instantaneous water heating system were almost twice as large.

4 COST COMPARISON

A detailed cost comparison between the two heating systems was done. Costs were estimated for:

- capital cost of feeder, service connections and meters
- losses in reticulation
- capital cost of water heater installed
- energy saving in IWHs relative to HWCs
- DSM saving by utility, possible only with HWCs

It was found that the instantaneous water heaters cost less from the consumers' viewpoint, but that for the supply authority the storage heaters were more cost effective.

The total cost over 15 years is lower for the storage heaters, illustrated in figure 5.

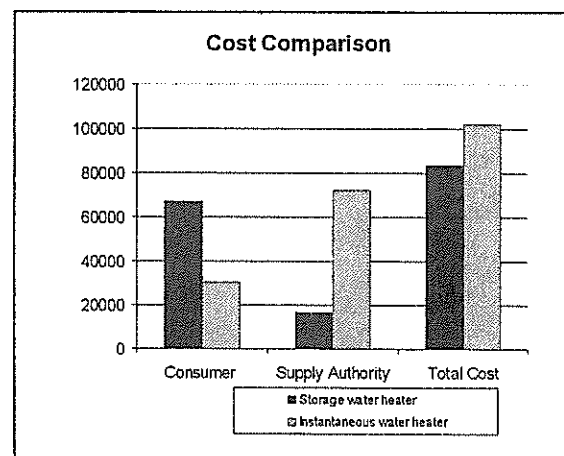


Figure 5: Cost Comparison

4 CONCLUSION

The cost of installing a reticulation system to supply consumers using IWHs is greater than the cost of the equivalent system with storage water heaters installed. The total cost of water heating with HWCs is lower than with IWHs.

Application of the Binomial Random Theorem also showed that for small groups of consumers, as on a feeder, the design current derived from the coincidence factors and consumer loads are not the same for the different types of water heating.

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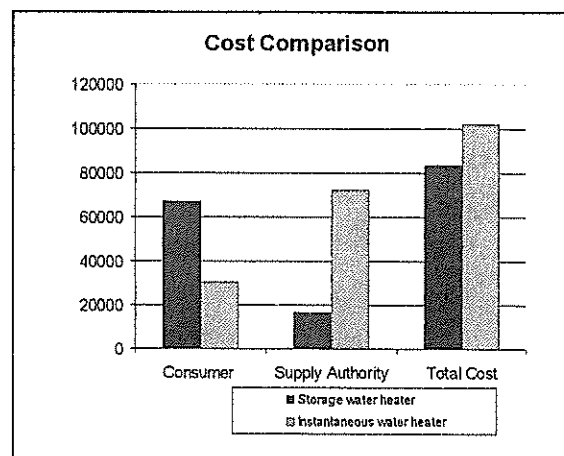


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Application of the Binomial Random Theorem also showed that for small groups of consumers, as on a feeder, the design current derived from the coincidence factors and consumer loads are not the same for the different types of water heating.

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