SIMPLE IMPROVED CIRCUIT MODEL FOR FUSE BREAKING TESTS

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1 Introduction

The breaking capacity of current-limiting fuses is verified using high-power test plant such as that shown in Fig.1(a), and the test conditions are conventionally characterised by five parameters for AC tests : the RMS test voltage, RMS prospective current, frequency, closing angle, and source-circuit power factor. From these quantities the simple equivalent circuit model shown in Fig.1(b) can be derived. This circuit model is implicit in the fuse test standards, and has often been used in the computer modelling of fuse breaking tests. This model will be referred to here as the *standard source-circuit model*. If the voltage generated by the fuse can be represented in some way, the current can be found by solving the differential equation :

$$\frac{di}{dt} = \frac{e - Ri - v_f}{L} \qquad (1)$$

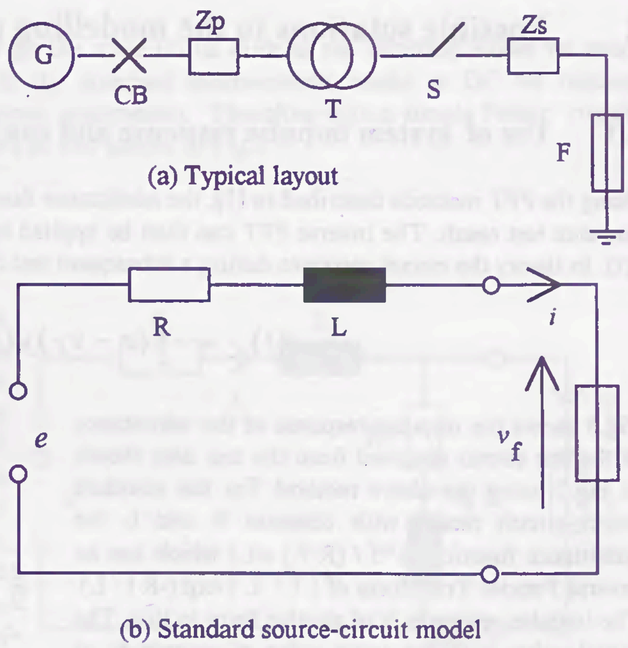

Fig.1 Circuit for testing current-limiting fuses

The values of R and L in the standard source-circuit model are evaluated from a calibration shot with the fuse replaced by a link, and they are the DC resistance and inductance of the circuit. However during the operation of current-limiting fuses the circuit current contains frequencies up to a few kHz, and use of the DC values can give a significant error.

Analysis more than 100 test records using Fast Fourier Transform (FFT) methods showed that the resistance in the model of Fig.1(b) increases strongly with frequency, with the value at 300Hz being up to 10 times the DC value [1]. This increase in resistance is due to eddy currents and skin effect in the various items of plant in the test circuit. Increases in resistance of the same order have been noted in other power applications [2, 3]. The effective series inductance L decreases slightly with frequency.

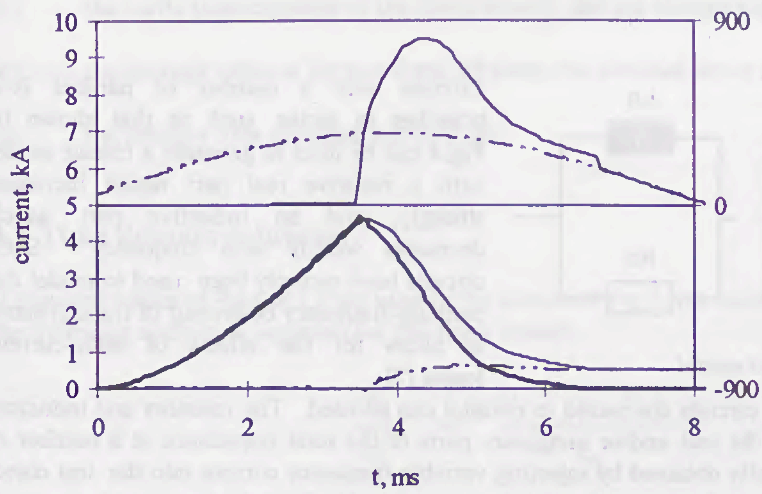

Fig.2 Error in current calculation using Perfect Fuse Model

The effect of this is on the circuit modelling problem is illustrated by **Fig.2**, which shows a comparison between test measurements and calculations using the 4th-order Runge-Kutta method. The upper traces show the measured fuse voltage with the known source e.m.f. The bold current curve is the test measurement, and alongside it is the current computed using equation (1) and a *perfect fuse model* (PFM). This means that that measured fuse voltage is used for the value v_f in (1).

If the standard source-circuit model is correct, and the perfect fuse model is used, then the computed current should agree with the measured value, but Fig.2 shows that there is an error. The error in current is also shown in Fig.2, growing rapidly at around 4ms.

The dynamic increase in circuit resistance during the test causes the circuit current to fall more rapidly than predicted by the standard source-circuit model, leading to an error averaging about +15% in the computed value of the I^2t integral.

This object of the work described in this paper was to find a better model for the test circuit, with a sound physical basis, but which is also simple to use.

2 Possible solutions to the modelling problem

2.1 Use of system impulse response and convolution

Using the FFT methods described in [1], the admittance function $Y(\omega)$of the source circuit can be derived from a one-shot test result. The inverse FFT can then be applied to $Y(\omega)$ to give the system impulse response function y(t). In theory the circuit response during a subsequent test can then computed using the convolution integral :

$$i(t) = \int (e - v_f) y(t - \tau)\, d\tau \tag{2}$$

Fig.3 shows the impulse response of the admittance of the test circuit obtained from the test data shown in Fig.2 using the above method. For the standard source-circuit model with constant R and L the admittance function is $1 / (R + j\,\omega L)$ which has an inverse Fourier Transform of $(1 / L) \exp(-R\,t / L)$. The impulse response is of similar form to this. The initial value is of the same order of magnitude as $(1 / L)$ but the response is multimodal with a fast initial decay followed by a slower one. An accurate expression for y(t) cannot be evaluated because of the presence of noise generated by inaccurate higher-frequency components in $Y(\omega)$ [1]. Furthermore, the convolution method cannot be very easily integrated with existing fuse models [4], so this method does not offer a practical solution.

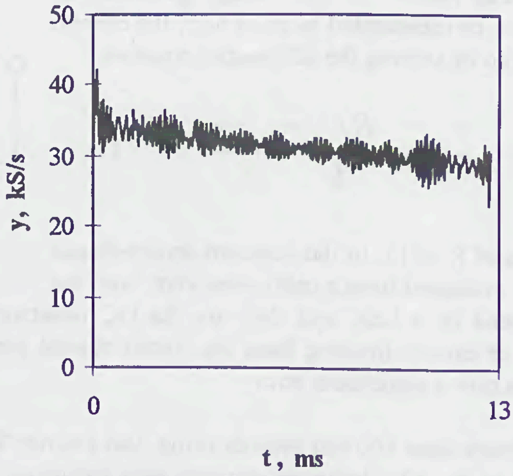

Fig.3 Impulse response of source admittance derived from test data shown in Fig.2

2.2 Series Foster circuits

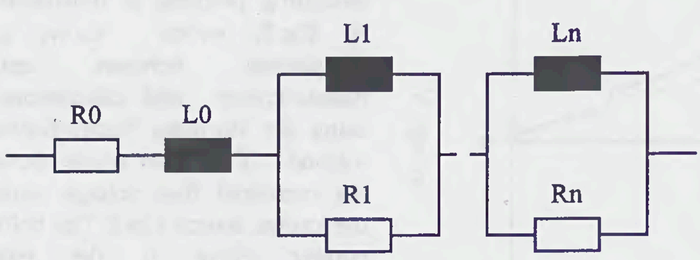

Fig.4 Foster series circuit model

Circuits with a number of parallel R-L branches in series, such as that shown in Fig.4 can be used to generate a circuit model with a resistive real part which increases strongly, and an inductive part which decreases weakly with frequency. Such circuits have recently been used to model the medium-frequency behaviour of transformers, to allow for the effects of eddy-current losses [2].

As an alternative a number of series R-L circuits connected in parallel can be used. The resistors and inductors in the model are found by matching up the real and/or imaginary parts of the total impedance at a number of discrete frequencies. Information is usually obtained by injecting variable-frequency current into the test object to determine the variation of impedance with frequency, but this is not practicable for a high-power short-circuit test laboratory, which has many different connections to give various prospective currents at several voltages and nominal power factors. We have only the impedance/frequency characteristics obtained from FFT analysis of test shots, which only gives data up to about 300Hz, and with considerable dispersion.

Using this data a number of low-order Foster-type alternative models have been investigated. It is easy to match up the increase of resistance with frequency to give more accuracy during the arcing period, but in each case the accuracy during the prearcing period was adversely affected, giving no worthwhile improvement in the total I^2t values.

Fig.2 shows that the standard source-circuit model gives good results during the prearcing period, and in particular the initial value of di/dt agrees well with the test data.

To match the high-frequency inductance L0, and to get the same initial di/dt as the standard model we need L = L0. However if we match the Foster model to the standard source-circuit model at DC we require L = L0 + L1 + L2 + ..., which conflicts with the previous requirement. Therefore with a simple Foster circuit we cannot compensate accurately for a discrepancy such as that shown in Fig.2.

3 Simple shunt correction circuit

Fig.2 shows that the error is quite small during the prearcing period but then grows rapidly when arcing begins, suggesting that the error is associated with the sudden growth in voltage across the fuse. During the calibration shot the supply transformer voltage and core flux will be very low, but during a test on a current-limiting fuse the transient voltage and associated losses will be much higher. This effect can be modelled by adding a shunt circuit such as that shown in Fig.5. During the prearcing period the fuse voltage and shunt current will be negligible, but as the fuse arc voltage rises the shunt current will grow in the same way as the error current .

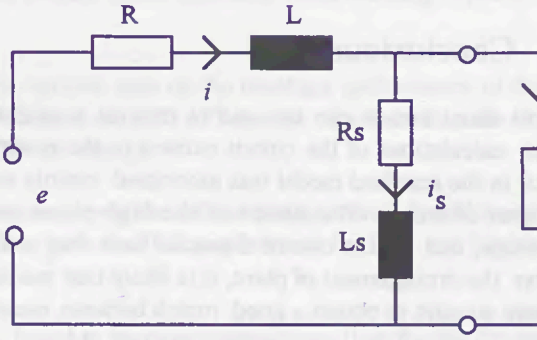

Fig.5 Simple improved circuit with shunt branch

The nature of a possible shunt branch was investigated by the following method. For each shot the fuse voltage and error current transients were transformed to the frequency domain using the FFT, giving the equivalent shunt impedance at each frequency. The shunt resistance and inductance were then evaluated from the real and imaginary parts of the shunt impedance. This analysis was carried out on 63 test shots on different fuses, with 5 different test circuit configurations, with the following results for the frequency range 0-300Hz:

(a) the equivalent shunt resistance decreased slowly with frequency

(b) the Ls/Rs time-constant of the shunt branch did not change significantly

(c) the average value of Rs was about 20 times the nominal series impedance

(d) the average time constant was 0.65 ms

3.1 Time domain solutions

If constant values of Rs and Ls are used in the circuit of Fig.5, the solution is easily obtained by solving (1) and the following additional equation for the shunt branch :

$$\frac{di_s}{dt} = \frac{v_f - R_s i_s}{L_s} \tag{3}$$

and the fuse current is then simply given by

$$i_f = i - i_s \tag{4}$$

The transients for all 63 shots were recomputed using the PFM and the circuit of Fig.5, using the values of Rs and Ls at 156.25 Hz obtained by FFT analysis of each individual shot. The error in each shot was almost completely eliminated.

However in a real situation we are not able to analyse a shot before it has happened, and we need to choose Rs and Ls somehow. With Rs set at 20 times the nominal source impedance and Ls set at Rsτs, with τs = 0.65 ms acceptable results were obtained for all data sets. The improvement over the standard source-circuit model is given in the table below.

	Mean error in computed I^2t	Standard deviation of the error
Standard circuit model	13.3%	5.96%
With shunt correction circuit	0.002%	4.82%

4 Conclusions

A simple shunt circuit can be used to provide a useful correction to the standard source model to get more accurate calculations of the circuit current in the modelling of fuse breaking tests. In the test results analysed the error in the standard model was associated mainly with shunt losses, which increase as the fuse voltage rises at the start of arcing. The nature of the high-power test laboratory plant and its connection varies between installations, and so the circuit discussed here may not be the most appropriate for other situations. However whatever the arrangement of plant, it is likely that medium frequency effects in the source circuit will need to be taken into account to obtain a good match between measured and computed transients.

5 Acknowledgement

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6 References

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