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A Multifrequency Phase-Coherent Pulse NMR Spectrometer: Description and Operation

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A Multifrequency Phase-Coherent Pulse NMR Spectrometer: Description and Operation

by

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A MULTIFREQUENCY PHASE-COHERENT PULSE NMR SPECTROMETER:
DESCRIPTION AND OPERATION

by
L. J. Burdett

ABSTRACT

A versatile multichannel multifrequency pulse nuclear magnetic resonance (NMR) spectrometer constructed for use at Los Alamos Scientific Laboratory is described. The spectrometer operates on the frequencies 8, 15, and 30 MHz, and the rf transmitter can be readily modified to operate at 16, 24, 32, 45, 48, 60, 90, or 120 MHz. Detail of the operation of the instrument, including circuit interconnections for most common pulse NMR experiments, is presented. Complete circuit diagrams are also included, as well as a discussion of a number of possible future modifications to the instrument.

1. INTRODUCTION

This report describes the design and operation of a pulse nuclear magnetic resonance (NMR) spectrometer constructed for use at Los Alamos Scientific Laboratory (LASL). Circuit interconnections for most common pulse NMR experiments are illustrated, as are some possible future modifications of the LASL instrument. Circuit diagrams are included in the appendixes.

In contrast to the more common continuous wave (CW) NMR experiments, the pulse NMR experiment begins with a disturbance of the equilibrium magnetization of the nuclear spin system. The return of the spin system to equilibrium, under various conditions, is monitored and makes possible the determination of a number of parameters of the resonance line. The most important parameters are the spin-lattice relaxation time T_1 , the spin-spin relaxation time T_2 , and the spin-lattice relaxation time in the rotating frame $T_{1\rho}$. If the dc polarizing magnetic field H_0 is considered to be in the z direction, then T_1 is the time constant for the return of the bulk magnetization to equilibrium along H_0 . T_2 is the time constant for equilibrium (dephasing) in

the x - y plane, and $T_{1\rho}$ is the time constant for establishing equilibrium along a field H_1 , rotating at the Larmor frequency in the x - y plane. Depending upon the particular system under study, these parameters may be equal or be considerably different with $T_1 > T_{1\rho} > T_2$.

Measurement of the above quantities and related parameters can often yield a wealth of information about the physical system under study. A number of books have been published on the general techniques and theory of magnetic resonance,¹⁻⁴ including an introduction to pulse methods.⁵ A short, and by no means complete, listing of information available from pulse NMR investigations would include: (a) studies of chemical exchange;⁶⁻¹⁰ (b) determination of chemical reaction rate constants;¹¹ (c) studies of molecular motion in both liquids¹² and solids;¹³⁻¹⁵ (d) measurement of chemical shifts and spin-spin couplings;^{12,16,17} (e) studies of solvation shells and complexes;¹⁸ (f) measurement of melting points and impurities;¹⁹ (g) studies of electrolyte solutions;²⁰ (h) investigations of polymers^{21,22} and glasses;²³ (i) studies of the properties of and motion in gases;²⁴⁻²⁶ (j) determination of the structure²⁷ of and impurities²⁸ in solids;

and (k) studies of self-diffusion²⁹⁻³³ in both liquids and solids.

Pulse NMR techniques are obviously a valuable addition to the commonly used methods of CW NMR. In addition, pulse methods are advantageous in high-resolution NMR spectroscopy,³⁴ and their application to Fourier transform NMR combines the advantages of CW and pulse techniques.

II. DESCRIPTION AND OPERATION

A block diagram of the spectrometer is shown in Fig. 1. The spectrometer consists of a dc pulse generator section, an rf transmitter, a transmitter-sample coil-receiver matching network, and a receiver. The receiver consists of three separate sections, a low noise rf preamplifier, a wide-bandwidth high-gain rf amplifier, and a detector-filter. Each component is described below.

A. Pulse Generator

Representative schemes for generating the sequences of gating (dc) pulses are shown in Figs. 2, 3, and 4. The wiring diagram for a spin-echo measurement of T_1 is given in Fig. 2. The pulse sequence used is a $(90^\circ - \tau - 180^\circ) - t_1 - (90^\circ - \tau - 180^\circ)$ sequence which repeats with the frequency of the Tektronix type 162 repetition rate generator, i.e., $1/t_2$. For accurate measurements, t_2 is set such that $t_2 \gg T_1$ for the sample under investigation. To vary t_1 , the calibrated delay option on a Tektronix type 545 or similar oscilloscope is used, and provides a stable and reproducible variable t_1 delay and readily

Fig. 1. Block diagram of the spectrometer system.

Fig. 2. Pulse generator wiring for a spin-echo measurement of T_1 .

allows measurements of the spin-echo amplitude as a function of t_1 , $M_z(t_1)$, to be performed. To obtain the value of T_1 , the data are then fitted to the expression $[M_0 - M_z(t_1)]/M_0 = \exp[-t_1/T_1]$, where $M_0 = M_z(\infty)$.

The 90° pulse length is controlled by a Tektronix type 163, and the long pulse generator is set to trigger at the end of the 90° pulse. The length of the pulse from the long pulse generator is set to be much less than τ . The pulse triggers the Tektronix type 162 sawtooth generator which gives one sawtooth ramp. The ramp triggers the 180° pulse length control (a Tektronix type 163), and its length in time provides the τ delay between the 90° and 180° pulses. The outputs of the Tektronix type 163 pulse generators are amplified and mixed in the pulse mixer (Fig. 2), and are channeled to the appropriate gate inputs of the transmitter. The device also provides positive and negative spikes at the beginning and end of the 90° and 180° pulses for time measurement and triggering applications. Timer outputs 2 and 1 can be used to measure t_1 , and outputs 1 and 3 can be used to measure τ . Circuit diagrams of the long pulse generator and the pulse mixer are given in Apps. A and B.

A similar diagram, useful for both T_1 and T_2 measurements, is shown in Fig. 3. The first pulse in the sequence is labeled the 9° pulse and the following pulses are labeled φ° pulses. Such an arrangement provides $(90^\circ - \tau - 180^\circ)$, $(180^\circ - \tau - 90^\circ)$ and $[90^\circ - \tau - (180^\circ - 2\tau)]_n$ sequences (depending upon the settings of the controls) which are repeated after each time interval t_2 . The time interval is set by the Tektronix type 162 repetition rate generator

Fig. 3. Pulse generator wiring for T_1 , spin-echo and Carr-Purcell sequences.

and again adjusted so that $t_2 \gg T_1$ for the sample being investigated. The $(90^\circ - \tau - 180^\circ)$ sequence is useful for the observation of a spin echo and also for measurement of T_2 by the so-called Method A.³⁵ In Method A, the spin-echo amplitude as a function of τ , $M_z(\tau)$ is measured and fitted to the expression $[M_z(\tau)/M_z(0)] = \exp(-2\tau/T_2)$. The behavior of the spin echo is often obscured by the effects of molecular self-diffusion³⁵ and the method is generally not useful except in the study of systems where T_2 is short, i.e., several hundred milliseconds or less. The $(90^\circ - \tau - 180^\circ)$ sequence is useful, however, for the measurement of D , the self-diffusion coefficient, if a linear gradient G is impressed upon the dc magnetic field H_0 . This can be accomplished by passing a steady current through a pair of reversed Helmholtz coils attached to the pole faces of the magnet. The $M_z(\tau)$ follows the expression $[M_z(\tau)/M_0] = \exp(-2\tau/T_2 - \gamma^2 G^2 D \tau^3 / 3)$ and, by calibrating G from the known diffusion constant of water,³⁶ D can be found for the sample under investigation.

The $(180^\circ - \tau - 90^\circ)$ sequence, or the $(90^\circ - \tau - 90^\circ)$ sequence, can be used to determine T_1 by measuring the amplitude of the free induction decay following the second pulse as a function of τ , $M_z(\tau)$. The interval between sequences, t_2 , is also set so that $t_2 \gg T_1$. The data are fitted to the expression $[M_0 - M_z(\tau)/2M_0] = \exp(-\tau/T_1)$ for the $(180^\circ - \tau - 90^\circ)$ sequence and to $[M_0 - M_z(\tau)/M_0] = \exp(-\tau/T_1)$ for the $(90^\circ - \tau - 90^\circ)$ sequence, where $M_0 = M_z(\infty)$. In both sequences, the output of the long pulse generator is adjusted so that the pulse length is less than τ . One ramp results from the Tektronix type 162 τ delay and

there is only one ϕ° pulse. For a $(90^\circ - \tau - 180^\circ)$ sequence the θ° length, controlled by a Tektronix type 163, is adjusted to give a 90° pulse and the ϕ° pulse length, also controlled by a Tektronix type 163, is adjusted to give a 180° pulse. In the $(180^\circ - \tau - 90^\circ)$ sequence, the reverse procedure is used, i.e., $\theta = 180^\circ$ and $\phi = 90^\circ$. Outputs 1 and 3 from the pulse mixer can be used to measure τ .

The $[90^\circ - \tau - (180^\circ - 2\tau)]_n$ sequence can be used to measure T_2 and is commonly called the Carr-Purcell sequence.²⁹ A Carr-Purcell sequence with the Meiboom-Gill modification³⁷ results when the phase of the 90° rf pulse from the transmitter is shifted 90° with respect to the rf phase of the 180° pulses, as just described. In this application, $\theta = 90^\circ$, $\phi = 180^\circ$, and the length of the output of the long pulse generator is set to provide n ramps from the Tektronix 162 τ delay generator. The sequence can be repeated after a time $t_2 \gg T_1$, and the timer outputs of the pulse mixer are used to measure the times τ and 2τ for adjustment purposes. Effects of molecular self-diffusion are partially removed by this technique, and successive echo maxima are fitted to the expression $[M_z(\tau)/M_0] = \exp(-2\tau/T_2)$ to obtain T_2 .

The pulse generator hookup for measurements of $T_{1\rho}$ is shown in Fig. 4. The configuration produces a 90° pulse followed, after a $7\text{-}\mu\text{sec}$ interval, by a 'long' pulse of duration t_w which is phase-shifted by 90° with respect to the first pulse. The $7\text{-}\mu\text{sec}$ delay is built into the triggering circuit of the long pulse generator and measurements are taken of the free induction decay amplitude (following the long pulse) as a function of t_w . These measurements are fitted to the expression $[M_z(t_w)/M_0] = \exp(-t_w/T_{1\rho})$ to obtain the spin-lattice relaxation time in the rotating frame $T_{1\rho}$.

Fig. 4. Pulse generator wiring for a $T_{1\rho}$ measurement.

B. Transmitter

The transmitter provides rf pulses to the sample and can be operated at any duty cycle up to 100%, i.e., continuous operation. The power amplifier output stage operates from a regulated dc supply which minimizes droop of the rf pulses and can deliver over 350 W of rf power to a resistive load. Three rf phase channels incorporated into the design allow the phase of a given output pulse, with respect to the internal oscillator, to be selected at will.

The transmitter consists of two sections: a solid-state rf driver, which contains the rf oscillator and rf phase-shifting networks, and a vacuum tube rf power amplifier. Three position switches mounted on the chassis permit selection of 8-, 15-, or 30-MHz frequencies. In addition, the simple modifications described in App. C will allow the frequencies 16, 24, 32, 45, 48, 60, 90, and 120 MHz to be generated. Front panel controls are simple, consisting of an OFF-ON rotary switch, the rf amplifier plate-tuning capacitor, a power amplifier screen voltage control which allows the rf output power to be adjusted, and rf phase controls for the three rf channels. There is a gate input for the entire transmitter and separate gate inputs for each of the three rf channels. The dc pulses from the pulse generator are applied to the appropriate rf channel gate input to select the desired preset rf phase of the output pulse.

The relative rf phase shift between the different channels must be calibrated with a high frequency oscilloscope, and the results of a typical calibration at 8 MHz for channels 1 and 3 are shown in Fig. 5. A detailed circuit diagram of the transmitter section is given in App. C.

C. Receiver

The receiver section consists of three separate components: a low-noise preamplifier, a wide bandwidth rf amplifier, and a detector. The preamplifier is an Arenburg Type PA-620-B and is operated as a narrow bandwidth device. Tuning controls for the inputs and outputs are located on the top of the chassis. In general, the input must be retuned at each different frequency and for each different sample coil.

The wide bandwidth rf amplifier is an Arenburg type WA-600-B which has flat response from 5-65 MHz and a gain of up to 80 dB. Although this amplifier

Fig. 5. Relative rf phase shift at 8 MHz for transmitter channels 1 and 3.

provides fast recovery after the rf pulse, it is quite sensitive to stray rf signals throughout the 5-65-MHz spectrum. In an attempt to decrease the bandwidth, a tuned circuit was attached to the grid of one of the amplifier tubes through a front panel connection; however, this modification provided only slight improvement. Two additional stages of rf amplification are provided by two Hewlett-Packard type 460AR wide-band amplifiers which operate in the frequency range 0.1-120 MHz and give up to 40 dB additional gain.

The detector stage which follows the rf amplifier consists of a simple diode detector followed by a low pass filter (Fig. 6). The low pass filter is a two-section elliptic-function (Cauer-Chebyshev) filter with a cutoff frequency of 25 kHz and a termination resistance of 1 k Ω . A similar detector with a filter cutoff frequency of 10 kHz has also been constructed, but losses in the inductors cause a somewhat less than satisfactory performance.

Fig. 6. Circuit diagram of the detector and filter. All capacitances are in picofarads and all inductances are in microhenrys.

This type of detector requires only simple linearity calibration which can be performed by weakly coupling the output of a modulated signal generator to the sample coil and recording the receiver output voltage as a function of input voltage. The weak coupling can be effected by placing a coil adjacent to the sample coil and applying to it the signal generator output.

D. Matching Network

The matching network tunes the sample coil to both the transmitter output and receiver input, and the circuit diagram for the 8-MHz unit is shown in Fig. 7. The circuit is somewhat unconventional in that it employs series tuning of the sample coil rather than parallel tuning. The plate-tuning capacitor, in the transmitter, and the inductance L_L form an L-network with $Q \approx 40$ which matches the output impedance of the transmitter power amplifier to the $3\text{-}\Omega$ series-tuned sample circuit. Crossed diodes D_1 and D_2 isolate the transmitter from the sample coil before and after the rf pulse.

The series-tuned sample circuit consists of the load resistor R_L , the tuning capacitor C_S and the sample coil L_S , which is connected to the matching network by 15 in. of low-capacity RG-114AU coaxial cable. The receiver is loosely coupled to the sample coil through C_R and protected by crossed diodes D_3 and D_4 . Similar networks for 15 and 30 MHz can be constructed by scaling the values of the components shown in Fig. 7.

There are advantages and disadvantages to the series-tuned type of matching network. The principal advantage is that in experiments requiring high power and long pulses, such as some $T_{1\rho}$ measurements, the bulk of the rf transmitter power is dissipated outside the sample chamber in R_L . The principal disadvantage of the design is that sample coil tuning

Fig. 7. Circuit diagram of the 8-MHz matching network. All capacitances are in picofarads and all inductances are in microhenrys.

(adjustment of C_S) and impedance matching (adjustment of the plate-tuning capacitor and of L_L) are somewhat more difficult and critical as compared to a single parallel-tuned sample circuit. In both types of matching to a single sample coil it is difficult to optimize both transmitter and receiver tuning simultaneously.

E. Tuning

When the correct pulse sequence is selected for an experiment, the magnetic field and pulse lengths must be set to their correct values. The simple tune-up procedure described here should help to eliminate any difficulties.

Using a sample with a relatively short T_1 , e.g., glycerol or water for a proton resonance, set the θ° and ϕ° pulse lengths to a nominal value, say 10 μsec . Find the resonance by sweeping the magnetic field H_0 until a free induction decay is observed. Adjust the θ° pulse length such that $\theta^\circ \ll 90^\circ$, i.e., 1 or 2 μsec , and vary the magnetic field until a maximum is observed in the free induction decay following the first pulse. Then tune the matching network, final amplifier plate, and receiver input for a maximum in the signal.

After completion of these steps, there should be a broad maximum in the signal as a function of the dc magnetic field. Set the magnetic field to the center of the broad maximum and increase the θ° pulse length until a maximum in the signal is again observed. Note the length of this pulse, using the time-interval counter, and increase the θ° pulse length to double this value. If correct tuning has been achieved, $\theta^\circ = 180^\circ$, and the amplitude of the

free induction decay will be zero. Small adjustments in the magnetic field and pulse length are generally sufficient to ensure proper tuning.

A rough check on the final tuning is to set $\theta^\circ = 90^\circ$, a maximum in its corresponding free induction decay; $\theta^\circ = 180^\circ$, a minimum in its corresponding free induction decay; and $\tau \ll T_2$ for the sample under consideration. A nuclear spin echo³⁸ will be visible at a time τ after the 180° pulse and will be of roughly the same amplitude as the free induction decay following the 90° pulse.

III. MODIFICATIONS

A number of modifications could be effected in this system which should, depending upon the particular measurements to be taken, improve the spectrometer performance. Five such changes are outlined below, roughly in order of decreasing importance.

A. Phase Detector

The substitution of a properly designed phase detector in place of the simple diode detector will result in improved signal-to-noise ratio for weak signals and greatly improved linearity. For a carrier injection voltage that is 50 times the maximum signal voltage, the linearity of a diode phase detector circuit is better than 1%. Such a linearity is difficult to achieve in practice, however, and linearity to within several percent is generally tolerable and can be achieved by keeping the maximum signal voltage equal to about one-tenth the carrier voltage.

A typical transformer type phase detector circuit is shown in Fig. 8. The 2N5245 FET operates as an rf voltage amplifier, with the variable delay

line in the gate circuit controlling the rf phase shift. The tuned circuits in transformer T_1 are tuned to the operating frequency, and capacitor C_C is a small coupling capacitor, e.g., 5 pF at 8 MHz. The low pass filter could be identical to that shown in Fig. 6 but designed with a higher termination resistance R_T , e.g., 22 k Ω . A simple capacitance filter could also be used.

Certain linear integrated circuits can also serve as phase detectors, and a typical circuit employing an RCA CA3005 is shown in Fig. 9. The carrier input requires approximately 3 V peak-to-peak from the variable delay line output, and the signal voltage need only be 500 mV, or less, peak-to-peak. The device can be powered by two 6-V lantern batteries, and the low pass filter can be identical to that discussed above. A Hewlett-Packard type 10534A double balanced mixer was tried as a phase detector but the low level input signals required, coupled with a moderate conversion loss, produced output signals too small to be useful without additional stages of video amplification. Use of this device, therefore, does not solve the problem of phase detection in this spectrometer.

B. Receiver

A modification which would decrease somewhat the versatility of the instrument but which should significantly enhance its signal-to-noise ratio would be the replacement of the Arenburg and Hewlett-Packard wide-band amplifiers with narrow bandwidth

Fig. 8. Circuit diagram of a transformer type phase detector. All capacitances are in microfarads.

Fig. 9. Circuit diagram of an integrated circuit type phase detector. All capacitances are in microfarads.

intermediate frequency (IF) strips for each frequency. This would effectively filter out much of the external rf interference present throughout the building and also, through the decreased bandwidth, reduce noise internal to the receiver system. Ideally, the main component of the internal noise arises from the thermal noise of the sample circuit and can be reduced only by decreasing either the bandwidth of the system or the temperature of the sample coil. Suitable narrow bandwidth IF strips can be purchased commercially or can be constructed either from vacuum tubes or from integrated circuits such as the RCA CA3023, operated as tuned amplifiers.

C. Matching Network

In applications where the dissipation of the transmitter power in the sample chamber is not important, as in short duty cycle or low power $T_{1\rho}$ experiments, a more efficient matching network could be used. Two examples of alternative matching circuits are given in Figs. 10 and 11. Figure 10 shows a simple parallel-tuned circuit, using a single sample coil with loose coupling to the receiver. The sample coil is tuned by C_S and the plate-tuning capacitor in the transmitter.

A crossed coil matching circuit is shown in Fig. 11. One advantage of this circuit is that the receiver and transmitter circuitry can be optimized separately, but there is the disadvantage of a somewhat more complex sample chamber arrangement. The transmitter coil L_T , usually a Helmholtz pair, is tuned by C_T and by the plate-tuning

Fig. 10. Circuit diagram of a simple parallel-tuned matching network.

Fig. 11. Circuit diagram of a crossed-coil matching network.

capacitor in the transmitter, and is damped by the resistor R_T , which should fall in the range 1-10 $k\Omega$. The receiver coil L_R is similarly damped with resistor R_R , which matches the tuned circuit to the cascode input stage of the preamplifier, and is tuned by capacitor C_R . Crossed diodes are used to protect the receiver input.

Designs for a number of matching circuits for pulse NMR have appeared in the literature in recent years,^{39,40} but the most efficient single coil design, useful for short pulse applications, appears to be that by Lowe and Tarr.⁴¹ The long lengths of cable make this circuit practical only at the higher frequencies, i.e., above about 20 MHz, but matching characteristics and recovery times are quite good. With any of the circuits cited above, the addition of a coil damping circuit could be useful, especially in broad-line studies where recovery time is important. Designs for a number of these devices have appeared in the literature, the most important being those by Samuelson and Ailion,⁴² and Roeder, Rhodes, and Schmidt.⁴³

D. Miscellaneous

Two additional modifications could be made to improve the performance of the spectrometer. The first is the installation of 10-turn helipot for selected controls in the Tektronix type 162 and 163 pulse generators. Such a modification would facilitate timing adjustments in the pulse generator section and increase its stability. The most important controls to be replaced are the waveform duration control on the type 162 τ delay generator and the output pulse delay control on the type 163 180°-length generator.

A shielded sample probe would also be useful in improving receiver system rejection of external rf noise. If accurate temperature control of the sample is required, however, then the project becomes difficult and serious consideration must be

given to the electrical characteristics of the probe and matching circuit.

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APPENDIX A

The circuit diagram of the long pulse generator is shown in Figs. A-1 and A-2. The device consists of a 7- μ sec delay multivibrator which triggers an

Fig. A-1. Circuit diagram of the long pulse generator

adjustable multivibrator. The second multivibrator provides pulse lengths of any duration between 0.5 μ sec and 20 sec. The amplitude of the output pulse can be as great as 24 V and is controlled by the 5-k Ω potentiometer at the output.

Fig. A-2. Detail of the long pulse generator range switch and power supply.

APPENDIX B

Circuit diagrams of one channel of the pulse mixer are shown in Figs. B-1 through B-4. The other channel is identical. The input amplifier is driven by a 5-V pulse, and six outputs are provided. Outputs 1 and 2 are for gating the transmitter and give

pulses of up to 70 V in amplitude. The other four outputs give both positive and negative spikes at the beginning and end of each pulse, and the spikes can be used for timing and triggering applications.

Fig. B-1. Input amplifier of pulse mixer, channel 1.

Fig. B-2. Pulse output circuits of pulse mixer, channel 1.

Fig. B-3. Circuit diagram of the negative spike generator in channel 1 of the pulse mixer.

Fig. B-4. Circuit diagram of the positive spike generator in channel 1 of the pulse mixer.

APPENDIX C

This appendix contains the circuit diagrams of the pulse NMR transmitter and a short discussion of its operation. The solid-state rf driver, containing the 8-, 15-, and 30-MHz oscillators, a broad-band rf gate, and the phase shifting networks, is the heart of the transmitter, and its block diagram is shown in Fig. C-1. Although four phase shifters are shown, only three have been constructed so far. All stages of the rf driver are gated and the circuit diagrams of the various blocks in Fig. C-1 are given in Figs. C-2 through C-6. The driver provides rf phase shifts over most of the 360° range and gives approximately 35 V peak-to-peak rf output. Figure C-7 shows detail of the power supply for the rf driver.

Circuit diagrams of the four-stage vacuum tube power amplifier, which follows the rf driver, are shown in Figs. C-8 through C-11. To generate the additional frequencies of 16, 24, 32, 45, 60, 90, or 120 MHz, mentioned in Sec. IIB, the tuning of the second and third stages can be altered. That is, with the 8-MHz oscillator, the plate of the 6CL6 stage could be tuned to 16 MHz to give a 16-MHz output. Doubling or tripling in the 5763 stage which follows would give 32 or 48-MHz outputs, respec-

tively. Similar schemes can be envisioned for the 15- and 30-MHz oscillators.

Detail of the control switch for the transmitter is shown in Fig. C-12 and the control relay wiring is shown in Fig. C-13. The switch positions of S_1 in Fig. C-12 correspond to 1-OFF, 2-FILAMENTIS ON, 3-TUNE, 4-ON. In the tune position, the screen grid of the 8122 power amplifier tube is grounded to limit the maximum plate current.

Circuit diagrams of the low voltage power supplies are shown in Figs. C-14 and C-15. Figure C-14a shows the +240-V and +150-V plate and screen supply for the first three stages of the vacuum tube rf amplifier. Figure C-14b shows the -30 V bias supply for the 8122 power amplifier stage. Figure C-15 shows the circuit detail of the adjustable screen voltage supply for the 8122 stage.

Tables C-I through C-IV indicate wiring detail on the 8 and 11 pin sockets and plugs located on the back of the instrument and are included for convenience in servicing. The notation used is as follows: K3,6 means that there is a direct connection from the indicated pin to pin 6 of relay K3; $S_{1b}(w)$ means that there is a direct connection from the indicated pin to the wiper arm of the b wiper of switch S_1 .

TABLE C-I

WIRING DETAIL OF THE 8 PIN PLUG, P1

Pin No.	Wire to
1	Ground
2	Screen supply
3	+240 V
4	+150 V
5	Ground
6	6.3 V ac
7	13.5 V ac
8	13.5 V ac

TABLE C-II

WIRING DETAIL OF THE 11 PIN PLUG, P2

Pin No.	Wire to
1	$S_{1a}(w)$
2	$S_{1a}(2,3,4)$
3	$S_{1b}(w)$
4	$S_{1b}(3,4)$
5	-30 V
6	110 V ac
7	110 V ac
8	Kepco P.S., red
9	Kepco P.S., white
10	Kepco P.S., black
11	Open

TABLE C-III

WIRING DETAIL OF THE 8 PIN SOCKET, S1

Pin No.	Wire to
1	Ground
2	Screen supply
3	K3, 3
4	K3, 6
5	Ground
6	6.3 V ac
7	13.5 V ac
8	13.5 V ac

TABLE C-IV

WIRING DETAIL OF THE 11 PIN SOCKET, S2

Pin No.	Wire to
1	K2, 9
2	K2, 10
3	K3, 2
4	K2, 9
5	-30 V
6	M
7	N
8	Open
9	Open
10	Open
11	Open

Fig. C-1. Block diagram of the rf driver. The letters label electrical connections between the circuit blocks with the letter G denoting a gate pulse input.

Fig. C-3. Schematic of the gated phase splitter.

Fig. C-2. (a) Schematic of the rf gate and buffer. In all figures, the resistors are in ohms and are $\frac{1}{2}$ W. Inductance is given in microhenrys; capacitance values greater than one are in picofarads, and less than one are in microfarads. (b) Switching detail of L_1 . That for L_2 and L_3 is similar.

Fig. C-4. Circuit diagram of the gated phase shifter.

Fig. C-5. (a) Detail of the buffer amplifier connections. (b) Buffer amplifier schematic.

Fig. C-6. Schematic of the output amplifier.

Fig. C-7. Detail of the rf driver power supply.

Fig. C-8. Input stage of the vacuum tube rf amplifier.

Fig. C-9. Second stage of the gated rf power amplifier

(a)

(b)

Fig. C-14. (a) Low voltage plate and screen supply for the gated rf power amplifier.
(b) Bias supply for the gated rf power amplifier.

Fig. C-15. Screen voltage supply for the output stage of the gated rf power amplifier.