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FOREWORD

The Conference Records of the Power Modulator Symposium continue to be a prime reference for the power electronics community. The device and applications papers on repetitive pulse power at this meeting were some of the best in my recollection, which goes back some 28 years. The quality of the papers can be traced to the leadership of Scott Gilmour, Program Chairman, and his dedicated paper selection committee.

This year's program brought a few new features. A poster session was organized by Richard Dollinger and was very well attended. The discussions generated during the poster session lasted far longer than the allotted time. We included a computer swap for the exchange of home-grown software free of any copyright restrictions as part of the poster session. The thanks for organizing the computer swap session goes to Lillian Dilks. The concept was a success and will be expanded at the next Power Modulator Symposium.

The keynote address, "Space Power for SDI Missions—A Road Map for the Future," was presented by Dr. John Hammond, Director of SDIO's Directed Energy Office, who has been a contributor to this symposium in the past and a strong advocate for pulse power. Jack provided very valuable insight into the planning for the SDIO systems of the future; correcting and clarifying some information that had been recently published on the subject. The audience was appreciative of the suggestions on areas of modulator and pulsed power technology that need attention.

It is the intent of the Executive Committee to continue featuring the devices and applications that are unique to this community. This is probably one of the few world-wide conferences that include sessions on ignitrons, thyratrons, and solid-state switches. Radar modulators continue to be a mainstay for the meeting. New uses for this technology are emerging in the fields of transportation and in manufacturing technology.

Now that the prior conference records are available from the State University of New York at Buffalo (Scott Gilmour) in a hard-bound series, future conferences will provide hard-bound copies of the Conference Records for the attendees. The success of the meeting, the accommodations, the location, the service, and the local arrangements are the result of the highly skilled staff of Palisades Institute for Research Services, Inc. Appreciation is also extended to the IEEE, Advisory Group on Electron Devices, DoD, committee members, authors, and attendees for their many contributions. The next Conference will be held in San Diego, California on June 26–28, 1990 and will be chaired by Dr. Larry Luessen of the Naval Surface Warfare Center, Dalghren, Virginia. Comments and suggestions regarding future sessions should be directed to any of the committee members.

Stephen Levy
U.S. Army LABCOM, ET&DL
Symposium Chairman

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MM WAVE MAGNETRON MODULATOR DEVELOPMENTS

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INTRODUCTION

This paper describes two recent developments to produce integrated modulator/power supply packages for operation with mm-wave magnetrons.

In the growing area of mm-wave systems the magnetron is one of the most practical power sources available. It is capable of much higher peak and mean power than any semiconductor source, while being the most efficient and cost effective of the tube sources.

With modern and well designed electronic packages, the overall performance as a power rf source for a radar system can be entirely satisfactory for many applications.

SUMMARY

Two recent developments are discussed.

The first unit described is for a ranging radar application and is a fully solid state line type modulator designed to operate a 35 GHz 60 Kw peak output power magnetron at prf's of up to 8000 pps, with a fixed rf pulse width of 50 nS. The unit was required to be of low weight (10 Kg), small size (3000 cm³), with low timing and pulse width jitter (<1 nS), and free from excessive pulse amplitude modulation (<0.1%).

The second unit described has evolved from a study for a complete electronic package to operate a recently developed 95 GHz 3 Kw peak magnetron. The package, incorporating all the necessary electronic services, heater, HT, LT power supplies with control and protection functions, had to operate over a wide range of pulse widths from 10 nS to 500 nS with prf and duty cycle limits of 50,000 pps and 0.001 respectively. Extremely compact size (250 cm³) and low weight (1 Kg) for the complete tube and electronics package was also required.

For both units the design philosophy will be discussed, the techniques used in the realisation of both the electrical and mechanical designs described and results will be presented.

35 GHZ RANGING RADAR MODULATOR/POWER SUPPLY SYSTEM

The outline specification required of the unit is shown in Fig 1.

The relative merits of possible techniques for modulator applications are well known (ref 1) and need little further explanation.

Because of the requirement for a fixed pulse width, the need to attain a small size and low weight, and the level of peak power to be provided, a solid state modulator of the SCR charge transfer, magnetically switched discharge type was chosen. This type of modulator is very robust and is well documented by other workers (ref 2 and ref 3). However care is needed in the design to ensure that the jitter specification can be met; realisation of the small size and low weight also requires special attention.

The radar equipment into which the system is incorporated provides the trigger pulse generation requirements, reset bias supplies, and magnetron heater voltage appropriately scheduled for duty cycle. The main thrust of the development was thus in the power conditioning and modulator.

The outline diagram of the modulator is shown in Fig 2 and conveniently splits down into two subassemblies.

1. Modulator Switching Assembly

The first of these is the modulator switching assembly, whose primary function is to control and regulate the voltage to which the primary storage capacitor (C2) is charged on each pulse. The operating voltage of this capacitor was chosen to ensure that regulation could be maintained throughout the range of mains voltage and load conditions over which operation is required. To minimise size and weight the charging regulator is operated from a rectified capacitor input supply working directly off the mains with a nominal 310 volts output to eliminate the need for an intermediate transformer or power supply. By design the voltage required on the capacitor C2 is between 450 and 500 volts so if dc resonant recharging is used then a voltage somewhat in excess of that required would be obtained. Because this aiming voltage is higher than required, a method of regulation is incorporated whereby, when the correct voltage on C2 is reached the energy remaining in the charging choke (L1) is rapidly diverted back to the dc supply. In this design the rapid diversion is achieved by switching off the power fet (TR1) and the energy is fed back to the dc supply reservoir capacitor (C1) via the secondary winding on the charging choke.

The use of the power fet in the abovementioned manner for the regulation also leads readily to the implementation of the command charge function, which is invariably required to allow scr recovery following the modulator discharge. The use of command charge also gives an important benefit to the regulator performance. By triggering the command charge a predetermined time before discharge is required, the post regulation droop on the capacitor (C2), caused by the monitor resistor and semiconductor leakage, will be constant irrespective of variations in prf. This means that capacitor C2 voltage regulation can be held to much tighter tolerances even under conditions of non-uniform prf.

By the use of a current monitor in series with the charging path the abovementioned circuitry can be made to produce rapid switch off under fault current conditions, such as may occur if the scr failed to recover.

The primary storage capacitor (C2) requires a back swing circuit to bleed off any excess inverse voltage due to modulator magnetron mismatches. With modulators that operate at short pulses, where the stray capacity energy is a significant proportion of the pulse energy, the inverse voltage can be a considerable fraction of the initial forward voltage. To make the unit more efficient the back swing network has an avalanche diode (D4) in series with it so that this normal mismatch energy is not bled off. On the next charge cycle this energy is recovered during the regulation process. To reduce the voltage stress on the power fet, charging is carried out via the backswing load resistor (R1), which is bypassed for charging via the choke (L2). Usually series inductance degrades the performance of charging regulators, but this does not occur with this circuit since when the power fet is turned off the current trapped in the choke will be dissipated in the backswing resistor. In practice this is a very small power (<1%) and is of little consequence.

The proposed circuit therefore provides an ideal charging control system for the modulator, incorporating backswing and overload circuitry, fault protection and an excellent regulation and ripple reduction performance. The control circuitry also incorporates trigger protection, trigger amplification, monitor functions and soft start signal to the solid state relay (SSR).

The design of the major components in the charging circuit is straight-forward and is covered in texts relating to dc resonant recharging (ref 4). However, there is the additional consideration of the specification of the time of the recharge trigger to ensure that energy diversion in the choke is complete before the discharge pulse occurs. Appendix 1 gives a derivation of total time required to complete the recharge and regulation function, and from this the recharge trigger timing can then be specified.

Because of the modest voltage level, (<1000 volts), the use of on line rectification techniques and the high prf, the final embodiment of the unit looks very similar to a switched mode power supply.

2. Pulse Transformer Assembly

In practice the pulse transformer assembly incorporates the bulk of the main body of the modulator components. From a circuit description standpoint the SCR and the primary storage capacitor (C2) which are within the modulator switching assembly, operate in conjunction with this unit.

The outline circuit is almost identical with those described by other workers (ref 2 and ref 3). Briefly, the circuit operation, assuming all reactors are reset to their respective negative remanant values and the primary storage capacitor is charged to the required value, is such that when the SCR is triggered on, the energy in the primary capacitor (C2) is resonantly transferred to the pfn capacitors (C3-C4). At the end of this transfer SR2 saturates and allows the energy to flow from the pfn to the magnetron via the pulse transformer. The reactor SR1 primes the SCR to assist in reducing its switching losses, and its saturated inductance determines the time of the resonant transfer. The reactor SR3 acts as a tailbiter to ensure a rapid fall of voltage at the end of the pulse. The backswing network (D5-C5-R5) minimises post pulse oscillation that could have a detrimental effect on magnetron life and/or system short range noise performance.

The main area of development with this unit has been the method of optimising and analysing the component values and the subsequent mechanical realisation of the unit. In other designs (ref 2 and ref 3) it is usual to step up in the transformer T1 such that the pfn impedance is typical of that encountered in a thyratron switched unit allowing the design of the components to be developed using traditional methods. In this design the opposite approach has been taken. The pfn in this system works at a similar voltage level to that in the modulator switching assembly. This means that the only high voltage engineering required in the system is the output of the pulse transformer where it interconnects to the magnetron. The low voltage design means that much more compact structures can be obtained, voltage stress and breakdown risks can be reduced, and that a large proportion of the unit can be fabricated from more readily available components.

Another departure from traditional design techniques is that of the pfn and its influence on the pulse shape. With very short pulses, where a large proportion of the stored pfn energy is required to bring the magnetron up to its working voltage, the more usual type E network (ref 5) tends to produce a very non-rectangular pulse. It is also well known (ref 6), that when pulse transformers are used with magnetron loads, ripples on the top of the pulse are produced that are difficult to suppress. Experience has also shown that interconnections to the magnetron can also have a significant effect on the pulse shape for the very short pulses.

It has always been possible to generate a rectangular short pulse by applying tailbiting to a much longer pulse. This usually results in a much larger unit, since effectively more energy per pulse is handled with the attendant losses. The ideal solution would be to square up the typical imperfect irregular shaped short pulse.

To address these problems a computer model of pfn/saturable reactor switch/pulse transformer and magnetron interface was developed and the circuit of this model is shown in Fig 3. In the model, the pfn is presented as a simple two stage system, not involving mutual inductance, and positioned in such a manner that series inductance of the capacitors constitutes part of the required inductance. The model calculates the loss, current, or voltage in any network to assist in the specification of the relevant components. The magnetron was specified in terms of its starting voltage, dynamic slope resistance and stray capacity. Many of the small effects such as interconnecting inductance to the magnetron, were also included. The model proved very useful and allowed optimisation of the pfn to partly compensate for pulse top ripples.

Having established a satisfactory paper design the next major aspect of the development was to produce a mechanical design to realise the desired electrical performance. A low voltage structure has been considered, so the resulting primary inductance figures were very low (of the order of 10 nH) and the pulse transformer ratio was relatively high (1:60 step up). To obtain the desired electrical characteristics the main switching reactor, pfn's and pulse transformers were integrated into a specially produced coaxial jig, which is shown in Fig 4. The central rod of the jig forms a single turn saturable reactor (SR2) with a one turn primary on the pulse transformer. The pfn is fabricated from four identical simple strip line units, connected in parallel and fed to a combining board to complete the circuit. The design produces very repeatable results due to virtually all inductive components being determined by machining processes and the accuracy of strip line artwork, two very accurate and repeatable processes. The saturable reactors use a stack of 15 off 0.0005" 4-79 Permalloy cores (or mu-metal), which are specified for loss and saturation flux capability. Within a given unit small variations between cores tend to cancel out and the saturable reactor has been found to be very repeatable. The pulse transformer is wound on 4-79 Permalloy core, suitably insulated with a ptfe box and located on the reactor tube by machined ptfe collars. A bifilar winding on this transformer allows the heater current to be fed directly to the magnetron. A fully screened current monitor is fitted on the pair of output leads. An epoxy moulded end hat provides the high voltage connection to the magnetron and also provides mounting for the tailbiting saturable reactor. The remaining components are assembled on to the low voltage end of the coaxial jig. These include the scr priming reactor (SR1), the bias hold off choke (L5) in series with the reset supply to all the reactors, and the coupling transformer (T1) which provides the important function of isolation from the modulator switching assembly and also allows the transfer to the pfns to be precisely matched.

The subsequent unit, shown in Fig 5 with its cover removed, is contained within a tubular case which is easily sealed. The high space utilisation means that the volume of oil required is very low, permitting the use of a simple and small bellows system, able to handle the fluid expansion associated with a large working temperature range. Fluid filling was chosen in preference to a solid dielectric system, such as epoxy resin, because of its lower weight, better thermal performance, lower dielectric constant helping to minimise stray capacity, and much easier servicing. The modulator unit, as arranged in the manner described, is a relatively simple arrangement of passive components, where all the major heat generating and high voltage components are all in one subassembly.

The results in service have been most encouraging and a typical rf pulse shape is shown in Fig 6.

The design techniques and the practical methods used in manufacture have proved easy to adapt to other modulators using different magnetrons and with variations in the design requirements. One embodiment of the design is used by Marconi Command and Control Systems in their type 282 Ranging Radar.

95 GHZ MODULATOR MAGNETRON PACKAGE

This project was in the form of a study to address the needs of an electronic package to operate a magnetron which was under development to meet the requirement of a reliable, compact and economic source of power at 95 GHz. The magnetron development was undertaken by English Electric Valve Company (E.E.V.) in conjunction with a study by Marconi Radar Systems Limited (M.R.S.L.) into suitable techniques for the future miniaturisation of the associated electronic package. Fig 7 illustrates the overall target specification for the magnetron electronics package.

The decision was taken at the outset to develop a unit whith a very simple application philosophy. The unit would accept just two inputs, a source of prime power and a trigger pulse sequence representative of the required rf output format. Internal to the unit there would be "housekeeping" control and protection functions but these would not require external stimulus or output any data for processing.

A broad assessment of the power requirements of the unit is given in Fig 8. From this it is clear that, both as a pulse modulator and as an electronic system in general, the unit is not particularly excessive in its power demand. At these power levels a large variety of circuit techniques can be considered. Using the size and power consumption figures illustrated in Figs 7 and 8 respectively, the dissipation density is of the order of 2 watts/cu. in. and if the unit is contact cooled on one face only then the heat flow density is of the order of 5 watts/sq. in. These figures are compatible with the employment of power hybrid assembly techniques which will be necessary to meet the low size and weight.

For the pulse modulator the requirement for pulse width flexibility implies a modulator of the hard tube type or a solid state equivalent of it. Pulse width flexibility is a fundamental system requirement and so rules out line type modulators.

Modulator Type

The type of modulator system chosen has a significant bearing on the attendant circuitry and three possible alternatives were considered.

1. A Hard Tube Modulator

A hard tube modulator using a small planar grid triode was considered a definite possibility. The electrical specification was well within the capabilities of this type of tube. To operate a magnetron working at around 6 Kv however would necessitate a dc supply of around 7.5 Kv. Voltage clearances around the large number of components that would be subjected to this voltage would tend to make the unit just too large for the required outline. The unit would also require its own heater supply of a similar order to that required by the magnetron.

2. A Solid State Switch Assembly to Replace the Planar Triode.

Power fets have made great strides into switching applications and it seemed sensible to pursue the possibility of applying them in this application. The use of fets in a series string as a replacement for a hard tube has been reported on by other workers (ref 7).

To evaluate the switching performance of a selection of devices a simple test fixture was designed with the circuit as shown in Fig 9, and the mechanical arrangement as shown in Fig 10. The fixture was designed to evaluate a single device and to simulate the magnetron type load an avalanche diode arrangement was used. This was scaled in proportion to the voltage at which a single fet in the string would operate. A series string of ten devices was anticipated as being needed so the avalanche diode voltage and the slope resistance were made one tenth of that of the magnetron (400 volts and 15 ohms respectively). The stray capacity was made some ten times the magnetron capacity (80 pF). Results with a number of devices was encouraging and a photogrpah of the current obtained at a 6 amp level is shown in Fig 11. The results indicated that good switching performance could be obtained but that the form of package of the devices, the industry standard TO220, was one of the limitations to switching speed, a difficulty also reported by other workers (ref 8). The move to hybridization, however, which would be necessary to achieve the size requirements, would allow the problems of package and interconnecting inductance to be addressed. As stated earlier, the power requirements are quite modest and so the fets evaluated were the smaller chip size devices with lower capacities to minimise switching losses. The Motorola MTP1N100 and the Philips BUZ50A both gave very similar performance. A problem was encountered, however, with the requirements of the low level power required by the driver stage for the fet. The distribution of even moderate levels of power, as well as the trigger pulse, to up to ten isolated stages was certain to increase the bulk of the unit to unacceptable levels. In fact it became clear that the overall switch, if executed in this manner, would be larger and heavier than an equivalent planar triode. Due to the importance attached to the overall size the series fet approach was not pursued.

3. Solid State Switch Assembly Using Parallel Devices in Conjunction with a Step Up Pulse Transformer.

The use of a pulse transformer, although offering size advantages, does present a number of complications.

The use of a pulse transformer generally results in a limitation on the maximum pulse width available and sometimes produces poor pulse shape at very short pulse output.

Pulse transformers are not specifically a limitation in themselves, but it is their use with a magnetron that poses the problem. Any attempt to make a fully distributed transformer such as the transmission line type, (ref 9), which would reproduce a very short pulse, provided they are correctly impedance matched, would encounter the problem of matching the magnetron to it. There are two basic reasons for this. Firstly the magnetron is a high impedance device of around 1000 ohms and at this level it is almost impossible to engineer a structure with appropriate dimensions to match this value. Secondly, the lumped capacity of the magnetron is usually significant - even more so when the capacity of the heater supply is taken into account.

A prototype of a conventional pulse transformer was made and it was constructed in a similar manner to that outlined for the ranging radar described earlier, except that the predominant dielectric medium was compressed air rather than oil. Secondary referred leakage inductance and shunt capacity values of 2.5 uH and 6 pF, respectively inclusive of the inter-connection systems to the outside of the package, were obtained with a 1:30 step up ratio possible. This transformer was only capable of supporting a pulse of up to 100 nS and thus fell short of the target specification.

A simple computer model of a number of parallel fets driving into a pulse transformer was developed and this is shown in Fig 12, with a typical analysis from this model shown in Fig 13. The pulse shape is reasonable but the rise and fall times are slower than desired. The model confirmed that improvements in the switching speed of the fets does not improve pulse shape performance, indeed the rise time of the fets has to be controlled to avoid a poor magnetron current pulse shape.

Thus, although the full pulse width range and pulse shape parameters could not be obtained with the pulse transformer approach, it was established that the size and weight criteria could be met for the system. Due to the modest voltage of the system an almost full hybrid approach could be considered, although the practical realisation of the pulse transformer would require special attention.

As the prime purpose of the first phase of the study was to demonstrate the full technical performance of both the system and the magnetron, the planar triode option was built as an evaluation model. The internal control philosophy was developed in such a manner that it can be readily applied to either the parallel fet/pulse transformer model or planar triode system.

Proposed System

The basic block diagram of the system is shown in Fig 14. The prime power of 115 v, 400 Hz is fed in via an rfi filter, to a full wave 3 phase rectifier with a centre tapped capacitor input filter. This provides a nominal 300 v dc supply and two 150 v supplies. The 300 v dc is fed to the main HT+ supply which is in the form of a resonant flyback convertor. The resonant aspect is achieved by tuning the transformer winding so that turn on of the fet always occurs when the voltage is at a low value. The full load switching frequency is approximately 500 KHz and the secondary consists of 10 individual outputs connected in series at their dc outputs. The flyback topology was chosen since a single wound component would give prime supply isolation, voltage transformation and energy transfer. The power supply control works on the quantum regulation principle in that the reservoir capacitor voltage is compared with a reference voltage and if it is low then the convertor will run at its full operating frequency at a preset duty. When the correct voltage is reached the convertor operation is inhibited by the signal fed via the opto coupler link. The operating frequency of the converter is also controlled by the voltage to frequency convertor. This gives the system an improved measure of mean power protection since if the voltage at the output falls too low then this will turn down the frequency of the convertor and create a current limit/foldback condition. An instantaneous peak current trip circuit is incorporated on the primary of the flyback convertor to add further protection. The power supply uses a feedback of the magnetron peak current in its control loop which gives good stability against variations that occur with ageing or temperature. In a variation of this control system, used on the fet modulator, this control method was particularly effective in compensating for "on" resistance changes with temperature in the fets.

The trigger input to the system is fed via a slow run in circuit to the control system reference voltage. In the absence of triggers the reference voltage is such that the ht will run up to just below its normal value, then on the commencement of triggering run up to its correct value. This important factor avoids voltage surges on the ht and more importantly prevents current transients that would unnecessarily tax the cathode of the magnetron and planar triode.

The trigger system incorporates protection for excess input voltage (positive or negative), accidental dc connection and excess pulse width.

One important feature that had to be addressed was that of protection of the magnetron should an internal arc occur. All magnetrons are prone to arc and if steps are not taken to minimise the arc energy, damage to the tube may result. With a magnetron designed for use at 95 GHz the internal geometry is very small. It was established early in the development program that using a line type modulator, with its inherent fault current limiting, and providing the stray capacity was kept to a low value (<60 pF) then the tube would not be damaged during periods of arcing. To obtain fault current limiting with this unit, the output of the magnetron current monitor is fed to a high speed comparator. In the event of the magnetron arcing, the voltage due to the fault current will exceed the comparator threshold level, turning off the grid drive to the planar triode. The output capacity of the unit was specified with a very low budget of only 15 pF and in the event a value between 16 to 18 pF was achieved. This was inclusive of modulator output, planar triode, magnetron heater system, magnetron and all other strays. The design aim of the arc protection circuitry is to provide a measure of protection at least as good as the inherent protection in a line type modulator. Tests conducted indicate that this has been achieved.

The LT convertor is a fairly straightforward multiple output flyback convertor operating at 500 KHz and is based on one of the standard 3825 chips. Again the flyback approach was chosen because of its single wound component configuration.

Two heater supplies are required, one each for the planar triode and the magnetron. Both supplies are required to be regulated against supply voltage variations and the magnetron also requires its heater power to be scheduled down as modulator average power is increased.

To simplify the electronic design hf ac invertors are fed via suitable transformers directly to the tubes' heaters. A number of tests were carried out to establish cathode temperature and emission performance for different frequencies of the ac current and it was established that at frequencies up to 100 KHz no appreciable degradation occurred. To regulate the heater supplies, the transformer that couples the output to the tubes is used as a current transformer and a true rms voltage convertor is incorporated into the control circuit of a pwm controller circuit again based on a 3825 chip. The circuits worked satisfactorily and the transformers for coupling to the heaters are very compact, even the unit required for isolating the magnetron cathode pulse voltage. Scheduling of the magnetron heater for increasing mean power is accomplished by taking an overwind from the main ht invertor transformer, integrating the voltage pulses so derived and using this to turn down the reference voltage in the magnetron heater control circuitry thus providing for correct heater operation over the required range of input duty.

To realise the design the main effort was concentrated on the mechanical interface between the magnetron and the planar triode output section and the transformer rectifier unit for the main ht supply. Fig 15 shows a view of the mechanical layout which is made more compact than originally thought possible. Three factors were significant in the making of a compact design.

1. The Magnetron Peak Current Monitor Transformer was integrated directly on to the cathode sidearm of the tube. The alloy section shown on the lower right of Fig 15 is actually a tube fitment and the current transformer is integrated into it. Careful magnetic field checks and appropriate amendments to the pole pieces allowed very close positioning of the current transformer without significant interaction with either the magnetic circuit or the transformer.

2. The Main Output Capacitor was fixed directly to the planar triode anode. This was achieved by using a matrix of high voltage chip ceramic capacitors.

3. A compensated potentiometer with low circuit loading capacity and adequate frequency response for the charging voltage monitor was made quite compact by the use of standard high voltage resistors and a coaxial construction technique. This item can be seen at the centre front of Fig 15.

The main ht transformer rectifier unit was made by having ten individual rectified outputs connected in series. Each secondary consisted of two single layer windings, one on each core limb, series connected. The ten individual windings are wound side by side which gives a gradually increasing voltage gradient over the length of the unit. The technique resulted in very low static and dynamic capacitance and low leakage inductance which is the same value for each winding. The prototype transformer rectifier unit shown in Fig 15 was quite large but it is expected that this can be reduced since one limitation found was that of handling and winding very thin wire. With more specialised techniques to handle very fine wire, a smaller transformer could be constructed.

The simple go and return type frame to house the planar triode and all its associated components together with the magnetron heater transformer and the magnetron current monitor is shown in Fig 15. An unexpectedly small modulator package was produced. No attempt was made to miniaturise the electronic package at this stage. The circuits used were assessed for hybridization by the hybrid manufacturers to obtain projected size and weight figures.

The results obtained when operating the unit with a magnetron, have been most encouraging, with both magnetron and modulator meeting a large proportion of the technical performance stated in Fig 8. The pulse shape performance is particularly encouraging. Typical results with a short pulse setting are shown in Fig 16.

An important aspect to be considered is that of the size. If one is prepared to accept a limited pulse width range with slightly degraded pulse shape, replacing the planar triode with a fet hybrid/pulse transformer configuration results in a much smaller unit since the bulk of the circuitry will be at a low voltage level. In particular the storage capacitors, voltage monitor potentiometer and TRU will be very much simplified.

The study program completed to date suggests that using modern methods of design and manufacture a totally integrated package to provide a lightweight source of power at 95 GHz can be achieved. The planar triode modulator can meet all the technical aspects of the system and can be kept surprisingly small in volume, virtually meeting the specification. A fet package could be smaller but would in general have a restricted pulse width range, possibly a higher minimum pulse width and lower upper prf limit.

CONCLUSION

The result from these two developments confirm that a magnetron integrated with a suitably designed electronic package can be a most satisfactory source for mm wave system applications.

ACKNOWLEDGMENTS

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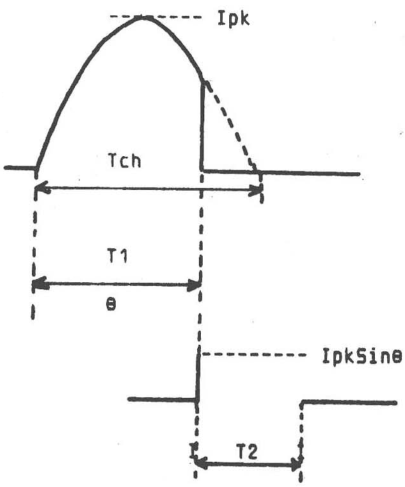
APPENDIX 1

Minimum time to complete charging process.

Assume a high Q and choke ratio of Unity (n = 1). When Fet turned on normal resonant recharging commences but with initial -ve voltage on capacitor. DC supply has a voltage of Edc and initial voltage on capacitor is -DEdc where D can be from 0 to 2.



If fet turned off at T1 and time for current to cease in choke is T2 then charging and regulation is fully completed in T1 + T2.

now $T1 = \frac{\theta}{\pi} \times \pi\sqrt{LC}$ where $Tch = \pi\sqrt{LC}$ normal recharge cycle under standard conditions.

also $Ipk = \frac{Edc(1 + D)}{z}$ where $z = \sqrt{L/C}$ the normal charging impedance

so $T2 = \frac{LIpkSin\theta}{Edc}$

substitute for Ipk and z so that

therefore $T2 = \frac{LEdc(1 + D)\, Sin\theta}{Edc\sqrt{L/C}}$

$$T2 = \sqrt{LC}\,(1 + D)\, Sin\theta$$

$$T2 = \frac{Tch}{\pi}(1 + D)\, Sin\theta$$

$$T1 + T2 = \frac{\theta Tch}{\pi} + \frac{Tch\,(1 + D)\, Sin\theta}{\pi}$$

$$T1 + T2 = Tch\left[\frac{\theta}{\pi} + \frac{(1 + D)\, Sin\theta}{\pi}\right] = Tch \times A$$

The solution of A for its highest value with respect to θ will thus give a multiplying factor over the normal resonant recharging time to allow a choice of pre trigger that ensures the recharge function is fully completed. For a range of D up to 0.5 typical results are given below.

D	A
0	1
0.1	1.01
0.2	1.024
0.3	1.043
0.4	1.065
0.5	1.088

REFERENCES

REF 1 Pulse Generators by G N Glasoe and J V Lebacqz MIT Radiation Laboratory series, Vol 5, Section 1.5.

REF 2 High Power Semi-conductor Magnetic Pulse Generators by G T Coate and L R Swain, MIT Press.

REF 3 Radar Handbook by M Skolnik, McGraw-Hill, Section 7.14.

REF 4 Pulse Generators by G N Glasoe and J V Lebacqz MIT Radiation Laboratory series, Vol 5, Chapter 9.

REF 5 Pulse Generators by G N Glasoe and J V Lebacqz MIT Radiation Laboratory series, Vol 5, Chapter 6.

REF 6 Pulse Generators by G N Glasoe and J V Lebacqz MIT Radiation Laboratory series, Vol 5, Pages 526-528.

REF 7 High Voltage Constant Current Solid State Modulator, G N Conrad, 17th Modulator Symposium, 1986.

REF 8 Power Mos Fast Switching Techniques, J A Oicles and G J Krauss, 16th Power Modulator Symposium, 1984.

REF 9 Nanosecond Pulse Transformers, C N Winningstad, IRE Transactions on Nuclear Science, March 1959.

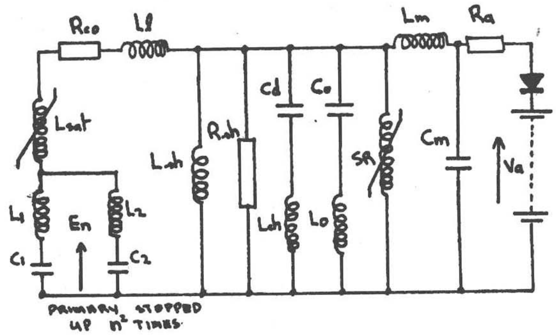
FIG1

RANGING RADAR MODULATOR POWER SUPPLY SPECIFICATION

Modulator Output Pulse	16Kv @ 16A
Magnetron Type	MG5329 (E.E.V.)
RF Pulse Width	50nS @ -3db
P.R.F.	8000pps Maximum
Timing Jitter	< 1nS rms
Anode Current Jitter	< 0.1%
Prime Power Source	230 volts +/- 10% 50/60 Hz Single phase
Temperature Range	-20°C to +60°C
Weight	10 Kg maximum
Volume	3000 cm³

FIG 3

RANGING RADAR MODULATOR ANALYTICAL CIRCUIT



PARAMETER	DESCRIPTION	VALUE
Va	MAGNETRON STARTING VOLTAGE	13.9Kv
Ra	MAGNETRON SLOPE RESISTANCE	100ohms
En	PFN INITIAL VOLTAGE	450V
L1	PFN INDUCTOR	6nH
L2	PFN INDUCTOR	1nH
C1	PFN CAPACITOR	0.15uF
C2	PFN CAPACITOR	0.04uF
Lsat	REACTOR SATURATED INDUCTANCE	9nH
n	PULSE TRANSFORMER RATIO	60
Ll	LEAKAGE INDUCTANCE (SEC REF)	10uH
Cd	TRANSFORMER CAPACITY (SEC REF)	28pF
Lch	Cd SERIES INDUCTANCE (SEC REF)	1uH
Rco	SERIES LOSS RESISTANCE (SEC REF)	40ohms
Rsh	SHUNT LOSS RESISTANCE (SEC REF)	45Kohms
Lsh	SHUNT INDUCTANCE (SEC REF)	9mH
Co	SHUNT CAPACITY (SEC REF)	15pF
Lo	SHUNT CAPACITY SERIES INDUCTANCE	0.3uH
Cm	MAGNETRON CAPACITY	10pF
Lm	MAGNETRON CONNECTION INDUCTANCE	0.4uH
Ls1	TAIL BITER INITIAL INDUCTANCE	1.2mH
Ls2	TAIL BITER SATURATED INDUCTANCE	30uH
VT	TAIL BITER VT PRODUCT	1.2V-mS
Jp	INITIAL PFN ENERGY	19mJ
J1	MAGNETRON INPUT ENERGY	12mJ
RRV	RATE OF RISE OF MAGNETRON VOLTAGE	290Kv/uS

FIG 2

RANGING RADAR OUTLINE CIRCUIT DIAGRAM

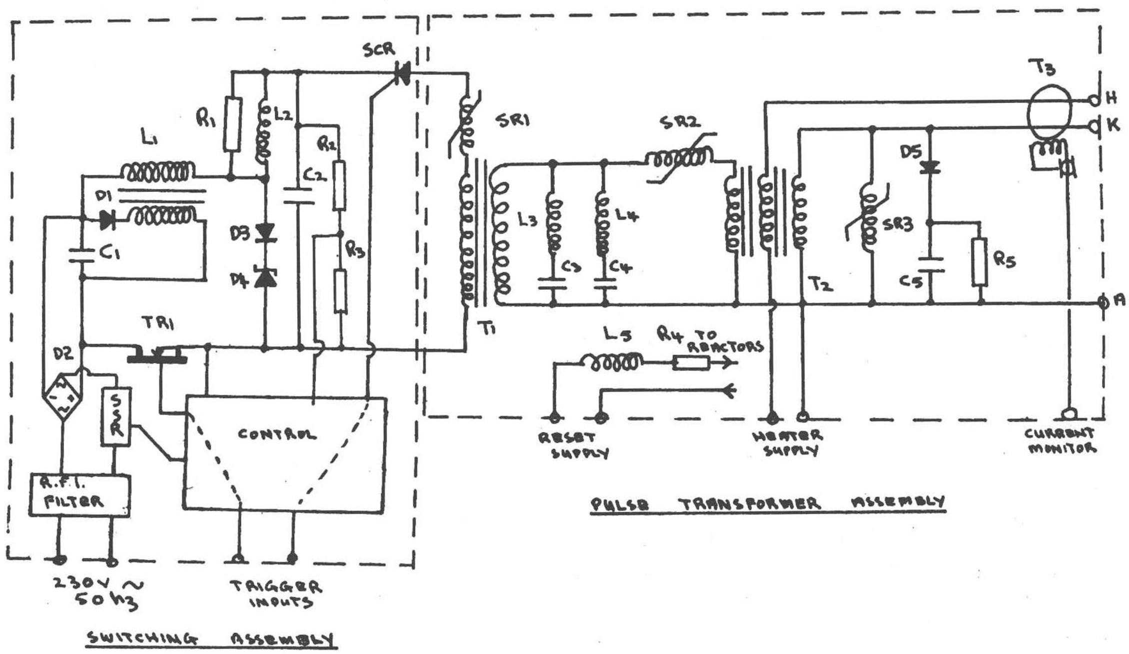


FIG4

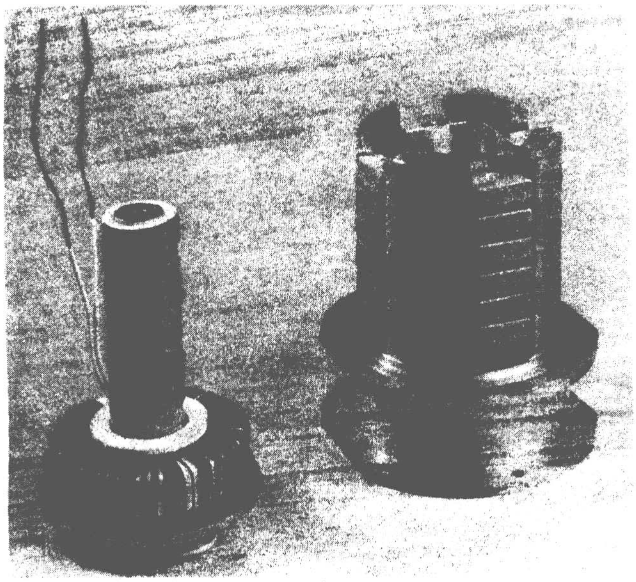
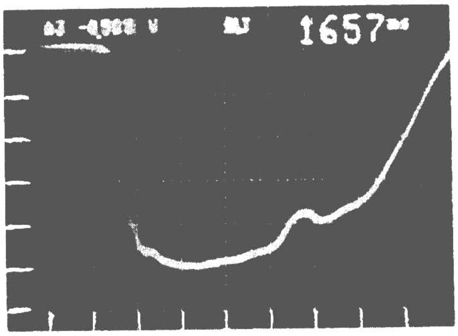


FIG 6

RANGING RADAR DETECTED RF PULSE



10nS/DIV

-1dB Width = 35nS
-3dB Width = 50nS
-7dB Width = 68nS

FIG5

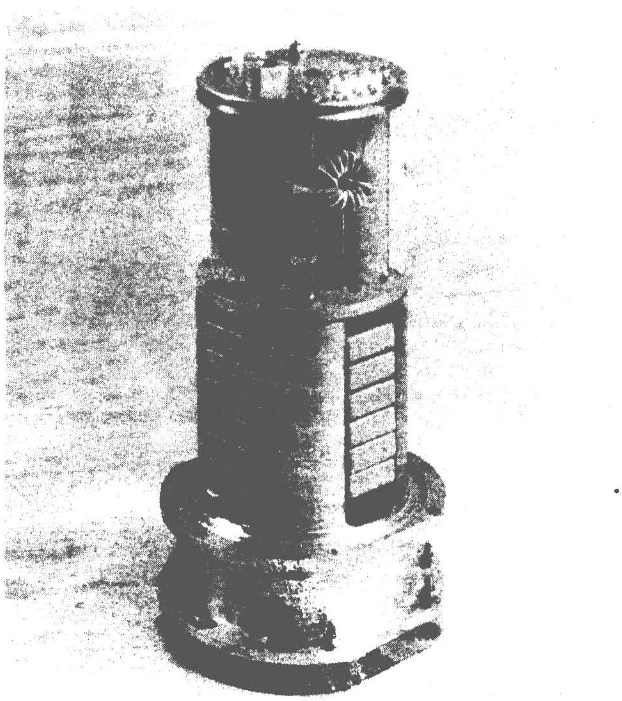


FIG 7

95Ghz INTEGRATED MODULATOR MAGNETRON PACKAGE

TARGET SPECIFICATION

RF Output	2Kw Peak 2W Mean
Pulse Width Range (continuously variable)	5ns Minimum 500nS Maximum
Rise Time (RF 10% to 90%)	< 3nS
Fall Time (RF 90% to 10%)	< 6nS
PRF	50,000pps Maximum
Duty Cycle	0.001 Maximum
Magnetron Pulse	6Kv @ 5A
Prime Power	115v 400Hz 3 Phase
Temperature Range	-20°C to +60°C
Target Volume	250 cm^3
Target Weight	1.5Kg

FIG 8

MODULATOR DESIGN - QUICK APPRAISAL

```
Tube input requirement (Va and Ia) = 6Kv at 5A
Duty cycle (du)                    = 0.001
prf maximum                        = 50,000pps

Tube capacity               = 6pF
Stray capacity              = 6pF
Heater circuit capacity     = 3pF

Gives a total capacity (Cd) of         15pF

At maximum prf this will give a power loss of

            Va².Cd.prf/2       =  13.5 Watts

Maximum magnetron input power is

            Va.Ia.du           =  30 Watts

Say modulator to psu efficiency = 80%

so dc power is (13.5+30)/0.8   =  54 Watts

Also tube heater could be     7 Watts
and low level power say       10 Watts

So total power is   54+7+10    =  71 Watts

Now assume prime power to
power required efficiency is 80%
so total power to package is

            71/0.8             =  89 Watts
```

FIG10

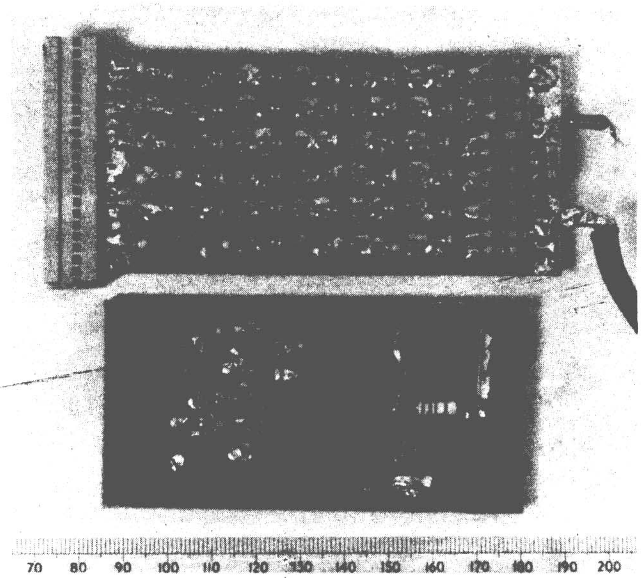


FIG 9

FET TEST MODULE CIRCUIT DIAGRAM

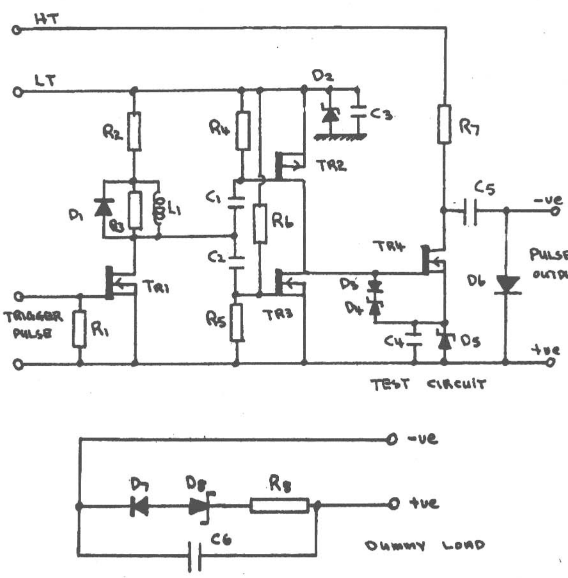
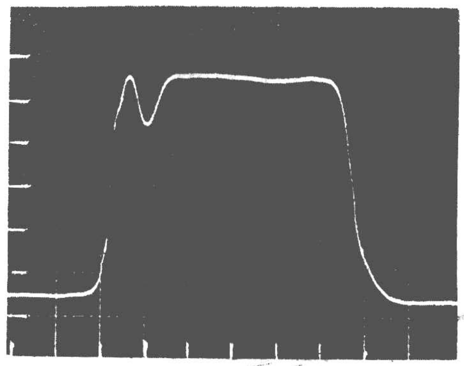


FIG 11

LOAD CURRENT FROM FET MODULE



10nS/DIV

6 AMPS PEAK