Innovations in Microwave Filters and Multiplexing Networks for Communications Satellite Systems

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***Abstract*—The advent of commercial satellite communications in the late 1960s provided a challenge to the microwave industry. It demanded technological advances to reduce the mass and volume of the communications payload and, at the same time, more efficient delivery of higher effective isotropic radiated power (EIRP) to the ground stations. One of the areas most dramatically impacted has been that of microwave filters and multiplexing networks. The past two decades have seen an order of magnitude improvement in the mass and volume of microwave filters. In the area of frequency combining or multiplexing, it is now possible to design and implement microwave multiplexing networks with arbitrary frequency spacing and channel bandwidths with an accuracy that was once reserved for individual microwave filters. This has allowed implementation of single mode transmit antennas. Such an antenna design provides maximum gain over a given coverage area and hence a maximum of EIRP. The net effect of these innovations has been the lowering of cost of a satellite channel. This paper describes the evolution of microwave filters and multiplexers for space application. The many advances are described within the context of the design of the overall communications subsystem. Extra emphasis is placed on the multipaction and passive inter-modulation (PIM) considerations in designing high power multiplexers. The impact of microwave filter technology in the channel characterization of satellite systems is described. The future directions of research and development are briefly discussed.**

I. Introduction

The advent of satellite communications is generally regarded as the launch of Intelsat I, II & III series of satellites in the late 1960s. These satellites established the viability of voice communications and retired many of the risks associated with satellite hardware. The launch of Intelsat IV in 1971—the first channelized transponder architecture—established the commercial viability of satellite communications. This was soon followed by the launch of Canadian domestic satellites and the upgraded Intelsat IV satellites referred to as Intelsat IV As. The frequency plan and a generic block diagram of the communications payloads of such non-regenerative satellites is described in Figs. 1 and 2.

In a communication system, the available frequency spectrum is a primary resource. The practical constraints of non-linearity in high power amplifiers (HPAs), necessitate channelization of the frequency band into a number

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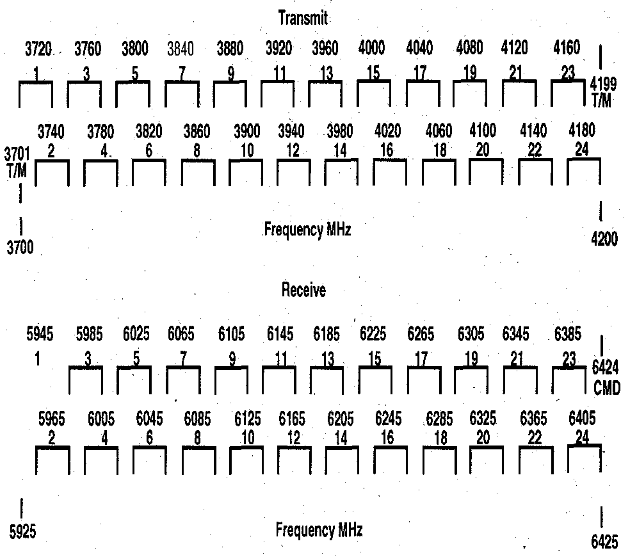

Fig. 1. Frequency plan of a typical *C*-band satellite.

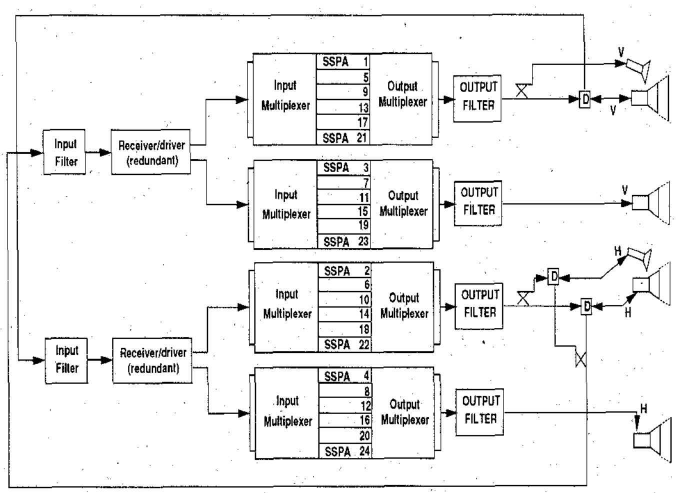

Fig. 2. Communications payload of a typical *C*-band satellite.

of RF channels, commonly referred to as transponders. Signals in each transponder are amplified separately and then re-combined at the output for transmission via a common antenna. This division into individual transponders and the subsequent recombining is accomplished by employing the input and output multiplexing networks (mux) in the communications payload. These multiplexers comprise RF channel filters coupled via ferrite circulators (input mux) or a waveguide manifold (output mux).

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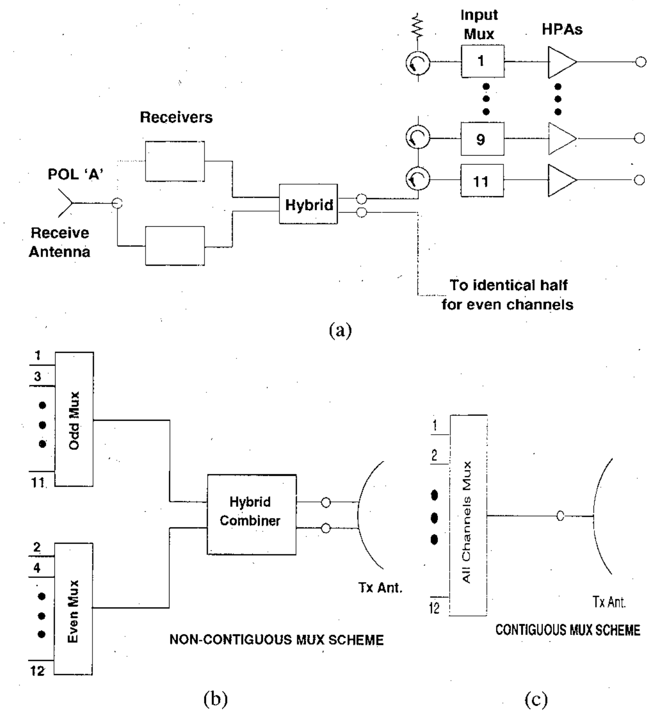


Fig. 3. Detailed block diagram of a typical communications payload. (a) Receive section. (b) Transmit section employing non-contiguous Mux scheme. (c) Transmit section employing continuous Mux schemes.

A more detailed diagram, describing the transponders on a single polarization is shown in Fig. 3.

The communication payload (Fig. 2) consists of receive and transmit antennas, low noise amplifier (LNA) and downconverter, input and output mux, and HPAs. The LNA and the downconverter is generally designed as a single cascaded unit referred to as the receiver. Antennas, receiver and HPAs are all relatively wideband devices and therefore have little impact on the amplitude and phase characteristics of transponders. Multiplexing networks, on the other hand, are the only equipment that deal with individual RF channels. The narrow-band RF channel filters in the multiplexing networks introduce transmission deviations, both amplitude and phase. Practical microwave filters require a certain transition band between the passband and stopband of the filter. This gives rise to tradeoffs between amplitude and phase dispersion in the passband and attenuation in the stopband, as a function of the guardband between adjacent transponders. The narrow bandwidth of these filters effectively governs the transponder channel characteristics; the wideband equipment contributing less than 5% to transmission deviations. As a consequence, the multiplex equipment has been a focus of attention, with the R&D push to optimize channel characteristics and, at the same time, to achieve a minimum of mass and volume for the multiplex equipment. For these reasons, microwave filter and multiplexing technology has gone through a series of innovations over the last two decades. This paper describes the historical developments, enabling technologies, current status and future directions of this important microwave equipment for space applications.

II. Filter Technology for Early Commercial Satcoms

The launch of Intelsat IV set a bench mark for the design and implementation of commercial communication satellites [1]. This system utilized the allocated 6/4 GHz frequency bands for the uplink and downlink respectively and, provided for twelve 36 MHz transponders on a single polarization within the allocated bandwidth of 500 MHz. Microwave filters employed for channelization used standard, rectangular waveguide structures based on classical filter synthesis techniques [2].

For the input mux, absolute insertion loss—as long as it is within a few dBs—is not a constraint. This is due to the fact that the noise figure is almost entirely governed by the front end receiver section and the following preamplifiers make up for the necessary gain in order to provide the desired input power to the HPAs. Any insertion loss after the Receiver contributes very little to the noise figure. As a consequence, the relatively lossy circulator coupled approach (Fig. 3(a)) is employed as it provides a maximum of flexibility, both in terms of realization as well as in the layout of the mux. In order to improve the group delay characteristics, input multiplexers also employ circulator coupled all-pass networks in cascade to achieve the required passband performance.

For the output section following the HPAs, absolute insertion loss is crucial as it directly impacts the EIRP. As a consequence, for the output multiplexing networks, signals in the non-adjacent channels are combined in a common waveguide manifold as shown in Fig. 3(b). Such an approach yields the minimum insertion loss. In this configuration, channel filters provide high isolation to signals in the non-adjacent channels, thus minimizing the adverse impact on the input impedances of those channels. This allows each channel filter to be designed on its own as if it were impervious to the presence of other channel filters on the common waveguide manifold. Implementation of such non-contiguous muxes using an empirical approach has proven adequate, although empirical designs do require extensive tuning to obtain optimum performance.

Multiplexing networks were implemented using standard .064 inches thick invar waveguides and standard flanges. As a consequence, the structures were rugged but massive. The average mass of channel filters per transponder was close to 4000 grams. This conservative design led to the first and somewhat obvious technique for mass reduction. The waveguide walls were reduced to less than .030 inches thickness, compatible with manufacturing processes of the time. This advance was incorporated in the filters used for the Canadian domestic satellite system launched in 1972.

Graphite Fiber Epoxy Composite (GFEC) Multiplexers

The first U.S. domestic satellites incorporated a number of advances, including the use of new materials for the microwave multiplexing networks. For narrow-band microwave filters, the waveguide material must be thermally stable, structurally stiff, lightweight and highly conductive. Materials like Kevlar, composed of light weight plastic fibers and used for aircraft structures met these criteria except for electrical conductivity. The challenge lay in depositing high conductivity silver onto the plastic fibers to make their performance equivalent to that of Invar waveguides. Manufacturing processes were successfully developed [3], [4] that emphasized, (i) the fabrication of graphite waveguide with a good internal surface finish and dimensional stability; (ii) the deposition of high conductivity plating, and (iii) the assembly of filter components into a design exhibiting frequency drift with temperature equivalent or better to that of Invar. This innovation was one of the key factors in the accommodation of 24 channels for the first time in the U.S. domestic satellite launched in 1975 [5]. Graphite filters are more difficult to manufacture than invar filters and require elaborate quality control and mechanical design. For these reasons and other advances made in the filter design and technology, GFEC filters are rarely employed in the present generation of spacecraft.

III. Dual Mode Filter Networks

At microwave frequencies, a filter resonator can be realized by enclosing a volume of space with a highly conducting material. Such a volume of space can support an infinity of electromagnetic field configurations or modes. Based on this premise, design of multi-mode filters, using degenerate modes with an identical natural frequency in a single cavity was first suggested by Lin [6] in 1951. However, to make a practical filter, it is necessary to suppress unwanted modes and be able to control, independently, the wanted modes. Lin's scheme did not satisfy these requirements. No one else took up this challenge and the effort was abandoned till the late 1960s when, with the advent of satellite communications, demands grew to make smaller and lighter waveguide equipment. In 1969, the first multi-mode cavity filter, a longitudinal dual-mode waveguide filter, was developed and later patented [7]. This early work at COMSAT laboratories established the viability of practical filters using more than one mode per waveguide cavity. Such a structure promised to reduce the number of physical cavities required for a filter by a factor of 2 with commensurate reduction in mass and volume. This promise coupled with the burgeoning demands of satellite industry resulted in unprecedented R&D activity in the area of multi-mode filter networks. Comsat Labs set the scene with a series of papers [8]–[11] establishing the theory and feasibility of a variety of dual-mode filters.

A typical dual-mode filter, realized in circular waveguides is described in Fig. 4. The basic principle is that a square or circular cavity has spatial symmetry. It can therefore support the dominant mode in either horizontal or vertical directions, thus providing two electrical resonant structures in a single waveguide cavity. This could be said of any mode that is not axially symmetrical. A

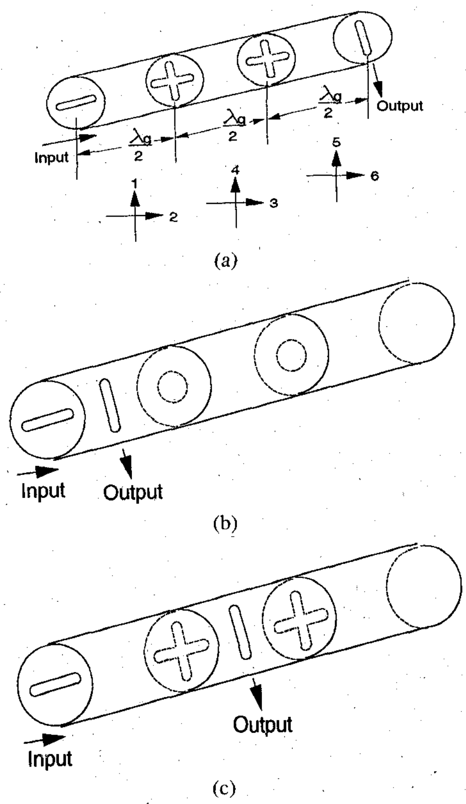


Fig. 4. Dual-mode filter realization in circular waveguide. (a) Longitudinal in-line structure. (b) Symmetrical canonical structure. (c) Asymmetrical canonical structure.

discontinuity at 45° angle between the two modes provides for a controlled coupling, yielding a practical filter structure. The only drawback, compared to the standard rectangular waveguide filters is that square or circular cavity resonators can support a higher number of spurious modes over wide bandwidths, making the design more complex. This is overcome by controlling the dimensions of the waveguides and irises and by optimizing the filter performance over desired bands of interest. The dual-mode structure provided another significant advantage. It allowed certain non-adjacent couplings between resonators which could be implemented simply by a cross iris or a coupling screw. This, in turn, allowed incorporation of transfer function zeros along the $j\omega$-axis or real axis or a combination thereof, thus permitting realization of a variety of response functions like elliptic, quazi-elliptic, linear phase etc. It therefore became possible to optimize the channel characteristics and hence the spectral efficiency without any increase in the mass. However, it did increase the design complexity. Prior to the dual mode waveguide structure, synthesis and realization of non-minimum phase response function like elliptic or linear phase was complex and involved large single mode folded structures [12], [13].

The fruits of this invention were first incorporated in the Intelsat IV A satellite launched in 1976 [14], the follow-on of Intelsat IV spacecraft. From then on, dual-mode networks became the industry standard for satellite applications.

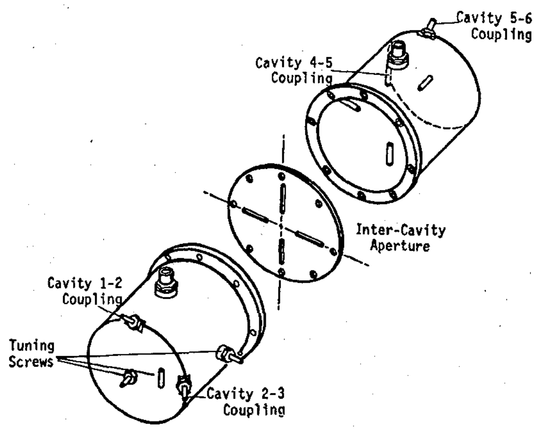


Fig. 5. Triple-mode structure with independent control of TE_{111} and TM_{010} modes.

Triple and Quadruple Mode Filters

After dual-mode filter networks, it was natural for researchers to focus their attention in realizing and controlling more than two modes in a single physical cavity. The dual-mode structure makes use of the spatial symmetry provided by the square or circular waveguides. The dominant resonant modes TE_{101} and T_{111} provide such a two-fold symmetry for the rectangular and circular waveguides, respectively. In order to have three modes in a single cavity, it was necessary to find another mode that could resonate at the same centre frequency and be orthogonal to the two TE_{101} or TE_{111} modes. Mode charts for rectangular and circular waveguides provide a variety of such modes [15]. Comsat Labs was the first to realize a triple-mode filter [9] by using two orthogonal TE_{111} modes and a TM_{010} mode to construct a two cavity six-pole filter. However, use of conventional slot or cruciform type irises for inter-cavity couplings (Fig. 4) prevented independent control of all the couplings simultaneously.

In the case of triple-mode filters, one requires three inter-cavity couplings between six electrical cavities in order to realize elliptic function filters. The coupling mechanism is provided by placing a metal plate with slots (iris) between the two cavities. However, the conventional irises used for dual mode structures can only provide a two dimensional independence for inter-cavity couplings. Therefore, in order to construct a two cavity six-pole elliptic filter, one must first find an iris structure that can control three inter-cavity simultaneously, such as described in Fig. 5 [16]. Triple-mode filters using such irises allow for independent control of TE_{111} and TM_{010} modes in cylindrical waveguides. Designs based on a cascade of dual and triple mode cavities are now being used for spacecraft multiplexers.

The quest for the quadruple mode filters required two orthogonal modes that could be used twice within square or cylindrical waveguide resonators. Such structures using TE_{113} and TM_{110} modes have been described [17], [18]. These structures are super sensitive and do not retain the performance over wide temperature ranges. To date, filters employing more than 3 modes per waveguide cavity have not been used for satellite applications.

IV. Evolution of Low Loss Filter Networks for 14/12 GHz Satellite Systems

The Intelsat IV and the Canadian domestic satellites established the commercial viability of the *C*-band (6/4 GHz) satellite systems. Expansion followed quickly with the launching of U.S., Indonesian and Brazilian domestic satellite systems. *C*-band Satcoms require large ground station antennas. This is partly due to the frequency band of such satellites and the restriction on the power that could be radiated from the satellite. The latter restriction is imposed to control interference with other competing ground based microwave communication systems. This led to the investigation of *Ku*-band (14/12 GHz) for satellite communications. This offered the advantage for large reduction in size of the ground stations antennas by virtue of the 3 fold increase in the operating frequencies; and, equally important, there is no restriction on the radiated power as parts of the band are exclusively allocated to the Fixed Satellite Service (FSS) in the International Telecommunications Union (ITU) region 2 which comprises North and South Americas. However, it did pose the technological challenge of equipment design at frequencies three times that used in the *C*-band satellite systems. The primary market for 14/12 GHz Satcoms was seen as Direct Broadcast Services (DBS) and business services at low cost. Therefore, maximization of EIRP emerged as a crucial issue to help achieve low cost earth terminals. This led to the higher power travelling wave tube amplifier (TWTA) development and spot beam antenna designs to allow beaming of higher power over selected portions of earth. It also posed a formidable challenge for designing microwave filter networks. The insertion loss in a waveguide filter depends directly upon the percentage bandwidth and inversely on the effective quality factor (Qu) of the resonant cavities. Qu depends on the skin depth and hence inversely on the square root of frequency. As a consequence, for *Ku*-band satellite systems, using standard rectangular or circular waveguide structures employing dominant modes, the insertion loss will be $3 * \sqrt{3}$ or 5.2 times that of *C*-band channels. This amounts to about 1 dB higher loss relative to equivalent 6/4 GHz transponders. It implies a potential loss of 20% in communications capacity for an equivalent transponder in *Ku*-band. In addition, high losses also degrade the in-band transmission characteristics resulting in further reduction of traffic capacity. The challenge was to develop filters at 12 GHz having a performance equivalent to those at 4 GHz. This could only be achieved by an increase in Q_u as nothing can be done about the increase in loss due to choice of frequencies. In practice, for a given resonant structure, it is possible to realize Q_u which is typically between 65 and 80% of the theoretical value. The only way to increase the value of Q_u in a significant way was to develop new types of resonant structures with inher-

TABLE I
EVOLUTION OF TE_{10n} AND TE_{11n} FILTERS

Year	Event
1975	Lab. demonstration of a TE_{103} dual-mode 12 GHz filter
1978	Canadian domestic satellite *Anik-B* launched First hybrid satcom using 6/4 and 14/12 GHz frequency bands First satellite to use low loss 12 GHz mux employing TE_{103} filters in square waveguide
1980s	Canadian Anik C series of satellites launched First satellite system to use low loss 12 GHz muxes employing TE_{113} filters in circular waveguide
1980s and Beyond	All satcom systems are using TE_{11n} or TE_{10n} ($n > 1$) low loss filters in the 14/12 and 30/20 GHz frequency bands

ently higher values of Q_u. The clue to meet such a challenge lay in a critical examination of the mode charts of rectangular and circular waveguides, particularly the theoretical quality factors associated with different modes [15]. The practical realization involved difficult tradeoffs in terms of achieving spurious free response over the entire up-link and down-link frequency bands for various higher order modes, constraints of minimum mass and volume and design simplicity that would allow control of response functions individually as well as in the multiplexed configuration. After various attempts, a new class of TE_{10n} and TE_{11n} ($n > 1$) filters were invented at COM DEV [19], [20]. The physical realization of these filters is similar to those described in Fig. 4 except that the cavity lengths are $n\lambda g/2$ where $n > 1$. This class of filters has the following advantages:

(i) Simple resonant structure, exhibiting a large increase in Q_u.
(ii) The physical structure lends itself readily to operate in single or dual mode configuration, thus providing a minimum mass and volume for a given Q_u.
(iii) The design allows a wide range of tradeoffs between the unloaded Q and the mass and volume of the microwave filter.
(iv) The physical structure is ideally suited to realize a wide range of filter functions in dual-mode configuration, including Chebyshev, elliptic, quazi-elliptic, linear phase, etc.

The foregoing advantages have allowed these types of filters to dominate the market for the *Ku*-band and higher frequency satellite systems for the past decade [21], [22]. The historic evolution of this class of filters for satellite systems is described in Table I.

V. Evolution of Contiguous Multiplexing for Satellite Systems

In the 1970s and early 1980s, the baseline design for multiplexer equipment was based on employing non-contiguous multiplexing schemes, as shown in Fig. 3(b). However, somewhere in the system, all channels had to be combined contiguously for transmission via a signal antenna. This, in effect, shifted the burden of contiguous combining to the antenna subsystem. In satellites, the contiguous combining of all channels is accomplished by feeding the outputs from the hybrid combiner to a common antenna feed network. This is commonly referred to as a dual-mode feed network. This is a simple scheme to implement but incurs a penalty in achievable antenna gain.

Fig. 6 describes a typical beam forming network employed for such a dual-mode antenna design. The two output ports of the hybrid contain all of the 12 channels at half the power. However, the odd and even channels at each port have a phase differential of 90°. This prevents realization of an optimum BFN, and requires a large number of components. Phase trimming must be carried out within all the network paths as well as between converters (couplers), and finally at the beam-former output ports. Even then only limited optimization of phase distribution over excitation ports is theoretically attainable, resulting in a penalty for the antenna gain. However, this scheme does allow implementation of the simpler conventional non-contiguous multiplexing scheme.

On the other hand, a contiguous multiplexing scheme, as shown in Fig. 7 allows for a simpler BFN, smaller parts count, phase and amplitude trimming to within close tolerances providing optimum antenna gain. However, implementation of this scheme requires that all channels be available at a single port i.e., a contiguous multiplexer.

The reason for using non-contiguous multiplexers for spacecraft in the 1970s has been simply that they could be realized practically whereas the progress in the design, and physical realization of contiguous multiplexers has been slow due to its complexity.

Relative to the conventional non-contiguous combiners, the design and implementation of contiguous muxes is more difficult. Theoretical modelling is complex and very sensitive to design tolerances. This is due to the larger number of channels (typically twice as many) on the same waveguide manifold in conjunction with large interaction between channels as they are adjacent to each other. This implies simultaneous optimization of a large number of variables. In addition, a thorough understand-

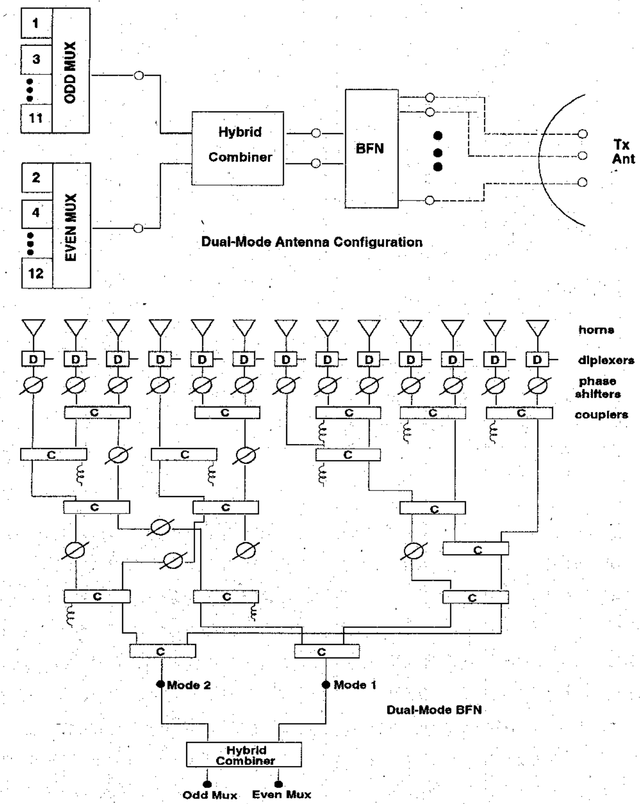


Fig. 6. Dual mode antenna configuration and BFN.

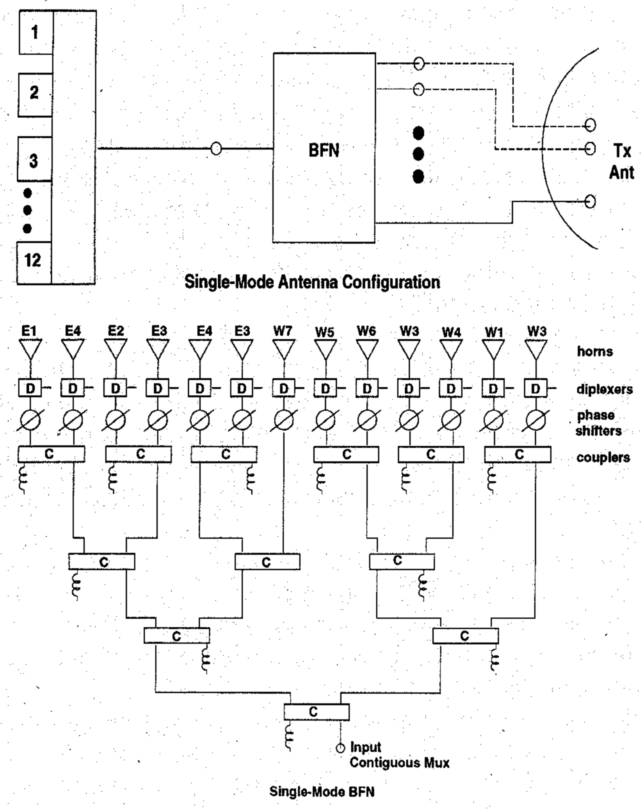


Fig. 7. Single mode antenna configuration and BFN.

ing of second order effects brought about by the manufacturing tolerances, discontinuities and the close proximity of tuning elements in the waveguide structure is required in the practical implementation of such multiplexers. Over the past few years, computer-aided optimization techniques augmented by extensive experimentation has brought contiguous multiplexer technology to the same level as the conventional non-contiguous design [4], [23]–[28]. This has been a difficult undertaking for many reasons. In a typical contiguous mux at *Ku*-band, there are eight channels. Each channel is typically composed of six pole channel filters realized in dual mode cavities and operating in the higher order low loss TE_{113} propagation mode. The desired response function is generaly quazi-elliptic requiring at least one cross-coupling in the filter structure. Each filter requires fine adjustments for its resonance frequencies, sequential couplings and the cross-coupling. For the six pole filter, this means a minimum of 15 variables including the input and output couplings. The manufacturing tolerances and system specifications necessitate very fine adjustments to meet the desired performance. When these filters are mounted on a common waveguide manifold, the spacing between channel filters and the location of the short circuit at one end of the manifold also become variable parameters. Thus, the total number of variables for optimization for an eight channel mux are 128. Such a large number of variables in conjunction with second order effects leads to an infinity of response shapes for the multiplexer. The sole use of a computer in optimizing such a large number of variables is not practical and perhaps not even feasible. What is generally required is a combination of optimization of a limited number of variables and practical experience in tuning multiplexers by observing the response shapes on an oscilloscope. Although it sounds like black art, this procedure by far has produced the best results to date. It is the authors' contention that attempts to tune contiguous mux using computer-aided techniques must involve some elements of pattern recognition for it is to be successful. Meanwhile, the experimental results on a wide variety of contiguous mux have clearly established its viability for satellites and ground stations. Measured performance tracks closely with computed values. Convincing satellite operators and system designers to adopt the more efficient contiguous multiplexing scheme for satellite payloads specifically involved [23]:

(i) Hardware proven software to demonstrate performance tradeoffs.
(ii) Demonstration of qualification model contiguous multiplexer.
(iii) Clear demonstration of enhanced EIRP via system tradeoffs and limited measured data.

Implementation of contiguous multiplexers for both the space and ground segments in operational systems is clear evidence of the success of this technology. The initial concepts for contiguous multiplexing were developed at COMSAT Labs. Ford Aerospace (now LORAL) was the first company to implement a 5-Channel *C*-band contiguous mux on the Intelsat V program. COM DEV was the first company to build high power *Ku*-band 8 and 12 Channel contiguous multiplexers for space and since then

TABLE II
EVOLUTION OF CONTIGUOUS MUX TECHNOLOGY

Year	Program	Multiplexer Hardware
1979	*Intelsat V*	5-Channel at 4 GHz
1983	*Intelsat VI*	5-Channel at 11 GHz
		5-Channel at 4 GHz
1984	*Satcom 'K'*	9-Channel, Dual-Tracking at 12 GHz
1985	*Intelsat VI (IBS + INT)**	8-Channel at 11/12 GHz (2 GHz BW)
1985	*Domestic Earth Stations*	High Power 6-Channel at 14 GHz
1986	*Superbird*	12-Channel at 12 GHz
1987	*Intelsat* Earth Stations	High Power 6-Channel at 6 GHz
1988		High Power 12-Channel at 6 GHz

*International Business and Telecommunications Services

has perfected this technology, both for space and ground segments. It is now possible to predict the performance, synthesize and realize contiguous mux with the same confidence and accuracy as non-contiguous mux or individual channel filters. Table II summarizes historical development of contiguous multiplexers.

ADVANTAGES OF EMPLOYING CONTIGUOUS MULTIPLEXING SCHEME FOR SATELLITE TRANSPONDERS

Contiguous multiplexers yield the following advantages relative to non-contiguous multiplexing:

Increased EIRP ($\cong 1$ dB) by allowing implementation of single-mode antennas;
Superior overall channel characteristics including multipath effects in the satellite;
Higher out-of-band isolation for suppressing signal spread from adjacent channels; and
A simpler, lower mass, single-mode antenna system.

VI. RESPONSE FUNCTION TRADEOFFS

An ideal filter response has no transmission deviation (flat amplitude and linear phase) in the passband and zero transition region between the pass and stop bands. Such an ideal response is not possible with a finite number of resonators of finite conductivity. However, advances in computer-aided design (CAD) and optimization techniques, coupled with dual- and triple-mode resonant structures, has allowed simulation and physical realization of filters with arbitrary amplitude and phase response. This advance has also benefited the process of computer-aided tuning of microwave filters and multiplexing networks.

In satellite systems it is desirable that all channels be symmetrical about their centre frequencies. It ensures maximum transmission capacity and no preference for carrier locations within a transponder. This is true for most transmission systems. Another constraint that is generally imposed from system considerations is the minimization of the transition region between pass and stop bands to maximize the usable bandwidth. This implies employment of equi-ripple class of function as they provide the sharpest cut-off regions. As a consequence, there is an abundance of literature in the analysis and synthesis techniques for the Chebyshev and elliptic function filters. However, there is always a trade-off available between pass and stopbands for a given order of filter. These trade-offs can be fully exploited using computer aided techniques.

For the class of equi-ripple functions with maximum number of peaks in the passband, it is most useful to employ the transformed variable technique [29] for analysing filter trade-offs. It is well suited for computer-aided analysis and optimization for this class of filters. This technique allows determination of the characteristic polynomials and hence poles and zeros for filters exhibiting the maximum number of equi-ripple peaks in the passband with arbitrary location (symmetrical) of the loss poles (transmission zeros along the $j\omega$-axis) or real-axis zeros. Elliptic and Chebyshev responses form a sub-set of filter functions that can be realized using this technique.

A further generalization was introduced [30] by treating characteristic polynomials independently without constraints. This permitted efficient computation of filter functions with arbitrary number of equi- or non equi-ripple peaks in the passband and arbitrary distribution (symmetrical) of zeros of the transfer function. This technique therefore provides a completely general treatment of filter functions with arbitrary amplitude and phase response, the only constraint being that the filter response is symmetrical. These functions are readily implemented in a dual-mode configuration.

Every so often, there is a requirement for asymmetric response filters. As an example, employing symmetric bandpass filters in a multiplexing network, the end channels have asymmetric response since there are no filters at the lower and upper edges of overall multiplexer bandwidth. One way to overcome this deficiency is to employ two extra filters, one at each end to render the response of end channels symmetrical. However, this incurs the mass and volume penalty of two extra channel filters. An alternate way to compensate is by employing asymmetric filters for the end channels. Often, suppression of specific spurious signals require asymmetric bandpass filters. The problem of asymmetric response function for equi-ripple class of filters was most effectively dealt by [31]. It allows generation of completely arbitrary asymmetric amplitude and phase response within the constraint of maximum number of equi-ripple peaks in the passband. Dual-mode realization of such asymmetric class of filters is well covered in [32]. Such filters are employed in satellite systems [33].

Allpass Equalizer Network

The response of the input mux channel filters often require group delay equalization for satellite application. This can be achieved by using non-minimum phase (linear phase) response functions for the channel filters. This entails the penalty of higher insertion loss, poorer isolation response, and a very sensitive structure. An alternative is to employ a circulator coupled all pass network operating in dual or triple-mode configuration. External equalization provides total independence between the filter and equalizer but incurs the penalty of the coupling circulator which adds mass and volume for the overall network. Based on tradeoffs for a large number of satellite systems [34], it is concluded that external equalization yields superior characteristics and is more cost effective to implement. The small weight penalty of the coupling circulator is often more than offset by the larger number of resonators required for the linear phase filters. It should be noted that the conclusion here is valid for typical satellite requirements and should not be extrapolated to imply universality. The coupling circulator for equalizer networks must exhibit extremely stable thermal characteristics to ensure close tracking between the response of the filter and equalizer networks. Over the years, the space industry has developed such stable ferrite devices. For these reasons, the bulk of the satellites in operation use external equalization for the input multiplexing networks.

VII. Multipath Effects in Satellite Channels and their Impact on Multiplexer Tradeoffs

In an operating satellite the incoming composite signal is channelized for separate amplification of signals in each channel to minimize intermodulation distortion as shown in Figs. 2 and 3. Signals pertaining to each channel are then re-combined and transmitted back to earth by a single antenna. This creates as many coherent paths for a signal as there are transponders in the satellite. Signal leakage through each of these paths add on to the main signal in a coherent manner, thus changing the total channel response. Contributions to amplitude and phase due to these leakage paths are referred to as multipath effects or contribution. In non-regenerative satellites, multipath effects are always present and significantly modify the overall amplitude and phase response of channels as observed by receiving ground stations. Over the years, it has become possible to compute and measure the multipath effects accurately [23], [35], [36]. Computer programs to simulate channel filters and multiplexers form an essential part of the software to simulate multipath effects. Based on many simulations and measurements [23], it is concluded that:

(i) In accordance with theoretical calculations, measured response of multipath contribution indicates that the optimum phase angle (as determined by coherent path lengths) for insertion loss, group delay and gain slope is not the same and hence it is not possible to optimize all of these parameters simultaneously.
(ii) Multipath effects are sensitive to any frequency offset between the input and output multiplexer networks for a given channel. Typical alignment tolerances of $\pm .3$ MHz can render the total loss and group delay variations asymmetric at band edges.
(iii) Multipath effects are also sensitive to the difference in gain between the adjacent transponders. A 2 dB difference which could be encountered in an operating satellite can alter the band-edge group delay and loss variations significantly.
(iv) When multipath effects are included for the channel characterization, there is often little to choose between the performance based on non-contiguous or the contiguous-band muxes.

Increasingly, multipath effects in non-regenerative satellites are now being included at the spacecraft level to ensure that the payload designs reflect optimum implementation of input and output multiplexing subsystems.

VIII. Reduced Mass and Size Using Dielectric Resonator Technology

It has long been recognized that the use of high permittivity dielectric materials offer large reductions in size and weight compared to conventional waveguide filters. However, only recent advances in material technology have made it possible to combine the high unloaded Q, good thermal stability and high dielectric constant in materials suitable for use at microwave frequencies. A ceramic compound $BaO{-}T_iO_2$ (barium titanate) was developed in the 1970s showing the desired mechanical, electrical and thermal properties. This spurred many researchers into action. Most of the early work was based on dielectric disk resonators to realize single-mode all-pole filters [37], [38]. Such filters offered little advantage over the dual-mode invar waveguide filters which were firmly established for space application in the late 1970s. In 1982, a dual-mode axially-mounted dielectric resonator loaded cavity filter was introduced on the scene [39] as shown in Fig. 8. It nearly matched the performance of conventional dual-mode waveguide filters and set the scene for the potential use of dielectric loaded multiplexers for space application. This structure is similar to the dual mode filters with the exception of loading the waveguide cavities with dielectric pucks thereby reducing their size. This simple extension of dual mode networks did incur a complex mechanical design. The dielectric pucks must be held tightly within the cylindrical cavities to withstand the launch environment and, at the same time, yield quality factor compatible with conventional structures. Nevertheless, this structure is being successfully used on a number of satellites. Subsequently, a planar dual-mode structure to realize dielectric loaded filters was invented [40], as described in Fig. 9. This planar mounted dual-mode dielectric resonator has good mechanical and thermal stability as well as being extremely flexible in the arrangement of cavities. This structure has also been successfully utilized in realizing triple-mode dielectric loaded resonator cavities [41]. This structure is emerging as the baseline design for the input mux of upcoming *C*-band satellites.

The two structures described here are suitable for input mux applications where the input power is typically -50 dBW. At such a low power, heat dissipation and conduction is not a constraint and is thus a secondary consideration in the design of dielectric loaded filters.

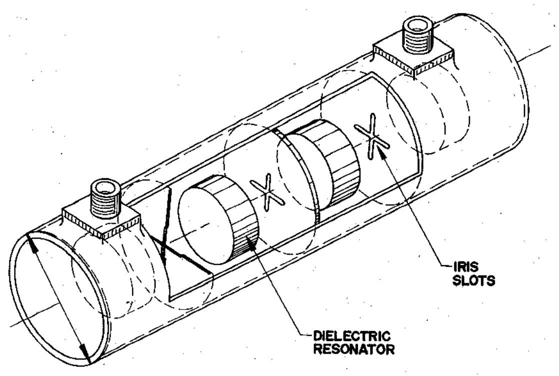


Fig. 8. Dual-mode dielectric resonator loaded cavity filter.

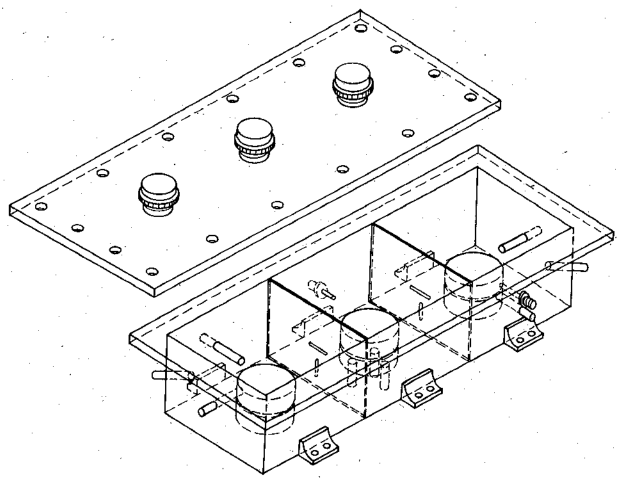


Fig. 9. Planar dual-mode dielectric resonator filter.

Power handling is limited by the amount of RF energy dissipated as heat in the dielectric resonators. The poor thermal conductivity of most dielectric materials coupled with the added burden of conducting this heat away from the dielectric resonators to the metal housing limits the thermal and hence the power handling capability of such filters. New types of dielectric filter designs have been introduced [42] which are capable of handling much higher power levels and yet provide a very compact structure.

The advanced configurations are based on $TE_{11\delta}$ and $HE_{11\delta}$ resonant modes of a cylindrical dielectric disk symmetrically positioned in a square cross-section cavity. The electromagnetic field patterns are relatively complex in such partially-loaded cavities but an attempt to illustrate their basic form is shown in Fig. 10.

The novel concepts involve taking advantage of the electromagnetic field structure of these resonance modes to create a quasi-multimode resonance regime within the cavity. This is done by introducing electrically-conducting surfaces within the cavity which intersect the dielectric disk, so splitting the normal resonating mode into several parts. Each of these modes resonates at the same frequency as before, but independently of each other until inter-coupled by some means. Thus the basis of a multi-element filter is formed, contained within the envelope of the original singly-resonant cavity. The metal walls penetrate to the inner regions of the dielectric resonators where the majority of the RF energy is dissipated, thus providing an efficient thermally-conductive path or a heat sink. This significantly increases the power handling capability of the resonator and extends their range of applicability to include high power (10 to 20 W/Channel) transmitter filters. Measured results on such structures [42] have established their feasibility for high power applications. Both of these advanced configurations operate on an image-resonator principle, as described in [43].

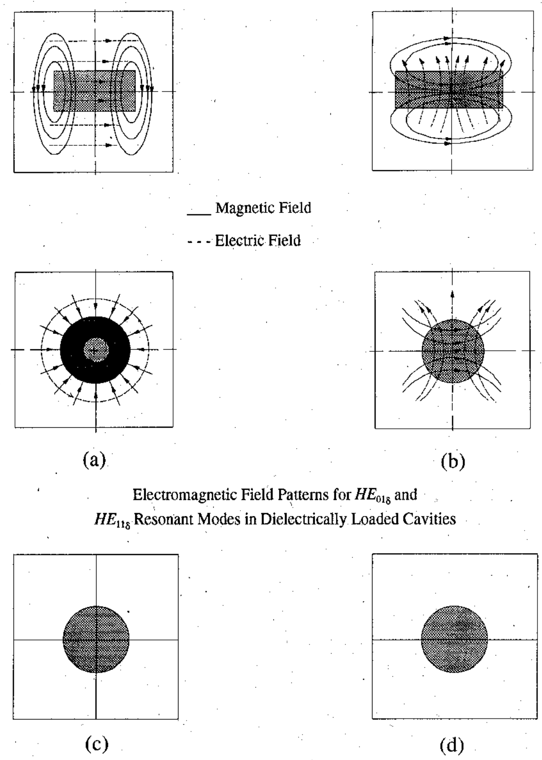


Fig. 10. Quarter wave cut resonator filters. (a) $TE_{01\delta}$. (b) $HE_{11\delta}$. (c) Two metallic dividing planes. (d) Single metallic dividing plane.

IX. Spectral Efficiency Criteria

The frequency plan of a communications system is defined by the centre frequencies and the available bandwidth for each RF channel. Of the available bandwidth, a small amount is apportioned as guardband between channels and the remainder is specified as the usable bandwidth. The ratio of this usable bandwidth to channel bandwidth is a measure of the utilization of the frequency spectrum. However, a simple statement that usable bandwidth shall be, say, X MHz, has no meaning unless it is defined in terms of measurable system parameters. In the past, there have been many definitions of bandwidths, ranging from customer stated bandwidth, Carson's Rule bandwidth [44], equi-ripple or return loss bandwidth, 3 dB or 1 dB bandwidths and so on. The ideal parameters would be the acceptable distortion levels due to channel imperfections for different types of traffic. This is difficult to evaluate and very expensive to demonstrate. A more practical way is to define the usable bandwidth in terms of the frequency response of a transponder. The fre-

TABLE III
CRITERIA FOR THE 'USABLE BANDWIDTH' OF A TRANSPONDER*

Parameter	Input (Rx Ant-HPA)	Output (HPA − Tx Ant)	Total (Input + Output + Multipath)
Gain Flatness, dB	<1	—	<2
Relative Group Delay, ns			
Middle 75%	<1.5	—	—
Over 100%	<35	—	<100
Isolation Protection, dB (Co-Polarized Channels)			
Middle 75%	≥35	>20	>55
Over 100%	≥25	>15	>40

*Simple and measurable parameters; accommodation of all types of traffic; frequency re-use with linear or circular polarizations; inclusion of multipath effects for channel optimization; 25 to 30 dB isolation between cross-polarized channels.

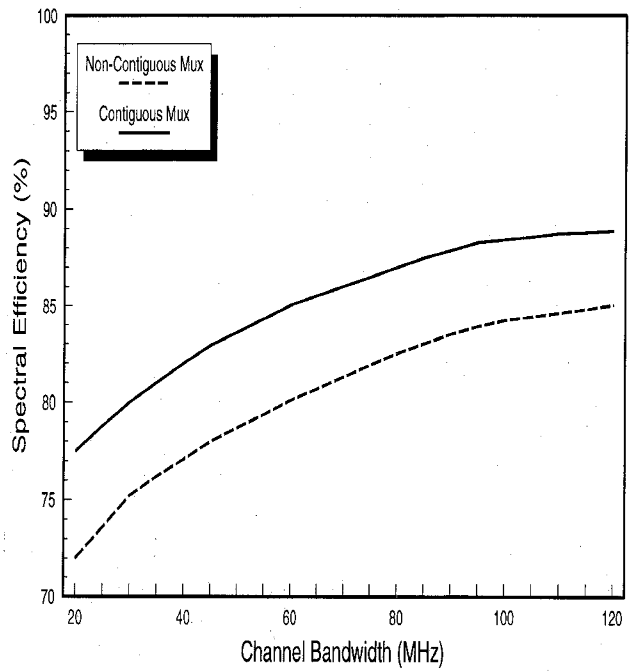


Fig. 11. Spectral efficiency of 14/12 GHz satellite systems.

quency response is primarily governed by the microwave multiplexing networks. Using this as a guide, a definition for usable bandwidth is proposed by including all effects in the satellite that influence the channel response and tempering it with state-of-the-art filter and multiplexer technology [45]. The key considerations in choosing these criteria and the corresponding parameters to define the 'Usable Bandwidth' are described in Table III.

As described in an earlier section, the advances in technology have made it possible to model and realize non-contiguous as well as contiguous multiplexers with a high degree of precision. This has allowed computation of channel response tradeoffs with multipath effects employing the two alternative multiplexing schemes and hence the evaluation of spectral efficiency using the criteria as described in Table III. Fig. 11 depicts the computed spectral efficiencies for 14/12 GHz satellite systems. These computations clearly show that use of contiguous multiplexers enhances the spectral efficiency by 5% relative to the conventional non-contiguous multiplexing scheme. Further improvements in spectral efficiencies are possible via improvements in materials (lower thermal expansion coupled with higher unloaded Q) or narrower operating temperature ranges. Future developments in these areas could push the spectral efficiencies closer to 90%.

Available frequency spectrum is a primary resource. The various regulatory bodies and satellite companies will continue to push to maximize the spectral efficiencies. It is reasonable to assume that developments in this area will continue in the future.

X. MULTIPACTOR BREAKDOWN AND PASSIVE INTER-MODULATION

Multipaction and passive intermodulation (PIM) continue to be a recurring problem in the design and implementation of high power microwave output mux and antenna feed networks for satellite applications. The reason for this is the complexity of these phenomena, their dependency on materials, assembly and workmanship, lack of adequate written material and the change of personnel every few years.

This section addresses the phenomena of multipaction and PIM as applied to high power microwave equipment for space applications. It includes discussion on the generation, detection and prevention of multipaction and a brief summary on PIM and ways to prevent it. It also includes test data for a range of microwave muxes reflecting the state-of-the-art, and concludes with an in-depth discussion on the adequacy (or lack of it) of design margins in specifying high power equipment for space applications.

Multipaction Phenomenon Description

Multipaction is an RF breakdown phenomenon. It takes place under conditions of vacuum when the mean free path of electrons is larger than the gap between the walls guiding the flow of RF power. The applied RF voltage will accelerate free electrons across the gap between the walls. Suppose that the transmit time is such that when the electrons hit the walls the field is reversed and the electrons are accelerated towards the opposite wall, thus creating an electron resonance. If the intensity of the applied field is such that electrons bombarding the walls are able to release secondary electrons continually, it can cause RF breakdown. It depends upon the electron multiplication via the secondary emission of electrons from the walls and is often referred to as the secondary electron resonance phenomenon. The generation of multiplication depends upon the following constraints:

(i) Vacuum condition.
(ii) Applied RF voltage.
(iii) Frequency of operation in conjunction with the geometry of RF components (fxd product).
(iv) Surface conditions.

A brief description of each of these constraints and its impact on the generation and sustenance of multiplication follows.

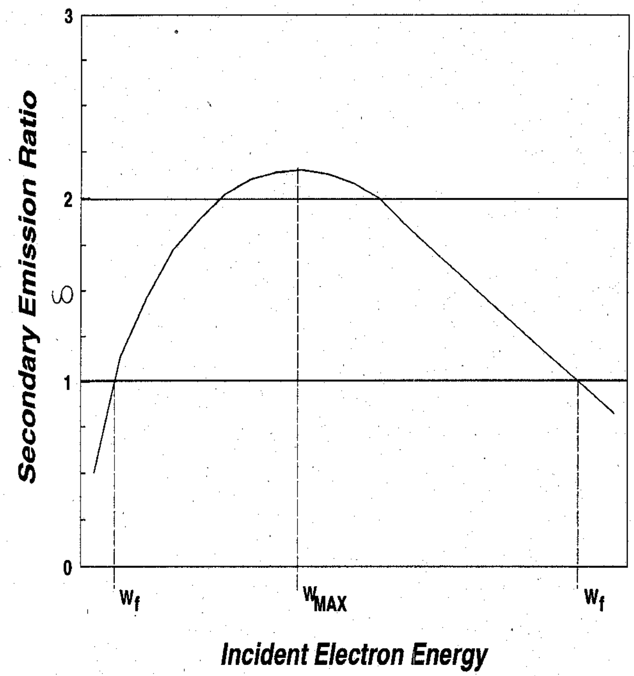
Dependence of Vacuum Environment

A condition for multipaction to occur is that the mean free path of electrons must be long enough to permit the electrons to be accelerated between the emitting surfaces with low probability of collision with ambient atoms or molecules. It has been shown [46] that for pressure of 10^{-3} torr or less, the mean free path of electrons is in the tens of cm range when surrounded by typical gas molecules like nitrogen, oxygen, helium etc. For RF equipment, typical gap sizes are in the millimeter range. It is therefore safe to assume that multipaction can occur under pressures of 10^{-3} torr or less.

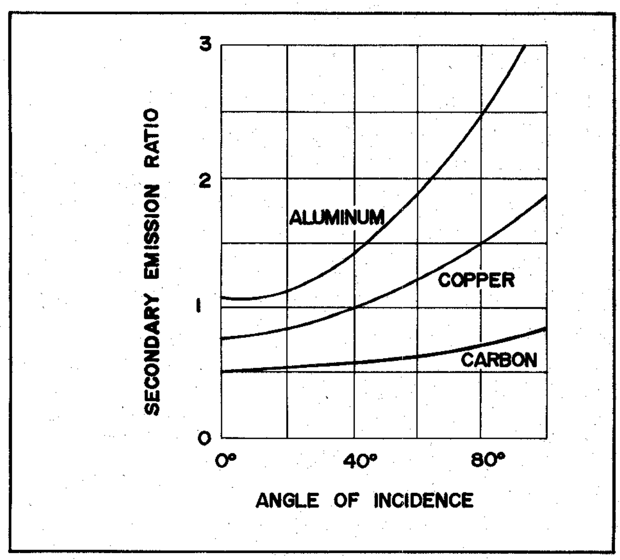
Dependence on Applied RF Voltage

The electrons are accelerated by the electric field created by the applied RF voltage. Under optimum conditions, the electron motion must be in phase with the field. However, the breakdown does not require the optimum conditions to occur, and there is a fairly broad region of fields and frequencies over which the phenomenon may be observed. A condition for multipaction is that the incident electrons on a wall must produce secondary electrons such that the ratio (δ) of the secondary to incident electrons is greater than one. If not, the generation of electrons will subside quickly and multipaction will not be sustained.

For a given material, the value of δ is a function of the energy and angle of incidence of the primary electrons [46] as shown in Fig. 12(a) and (b). At low incident energies, primary electrons are unable to liberate many secondary electrons and multipaction does not get initiated. At very high incident energy, the primary electrons penetrate so deeply into the surface that the secondary electrons produced are trapped in the substance and do not reach the surface. The energy of the primary electrons must lie between these minimum and maximum values to sustain multipaction breakdown as shown in Fig. 12(a). This energy depends upon the phase angle of electrons



(a)



(b)

Fig. 12. Secondary emission parameters. (a) δ versus W. (b) δ versus θ. (From NASA Report CR-488)

with respect to the applied RF field. As a consequence, there is a phase window within which the incident electrons may produce secondary electrons and hence multipaction as shown in Fig. 12(b). These upper and lower bounds on energy and phase permit multipaction to occur over a broad range of frequencies and applied voltages.

Dependence on $f \times d$ Product

A condition of multipaction is that the gap size correspond to a multiple of half-cycle of the applied RF voltage to satisfy the condition of electron resonance. The simplest geometry to determine breakdown as a function of $f \times d$ product is that of a parallel plate geometry. Many researchers [46]–[48] have made extensive measurements in such a way as to obtain both the maximum and minimum breakdown field strengths at various excitation frequencies. A typical depiction of this relationship is shown in Fig. 13. There is a multipaction curve for each odd multiple of the gap width as it corresponds to the half-cycle of the RF voltage. The primary mode corresponds to the multiple unity. Other modes are referred to as high order multipaction modes. Each multipaction curve has a minimum and maximum value of energy or peak voltage to initiate and sustain multipaction as described in the previous section. Also, each multipaction curve has a width which corresponds to the angles of arrival within which the secondary emission coefficient is greater than unity to sustain multipaction. Such curves are available for a number of pure materials.

The geometry of high power RF equipment rarely corresponds to simple parallel conductors. Discontinuities and complex geometries are generally required to meet electrical performance requirements. In the vicinity of discontinuities and sharp edges, the electric fields can be much higher than those based on simple parallel geometry. As a consequence, multipaction can occur at RF voltages seemingly lower than those based on parallel plate geometry as shown in Fig. 13.

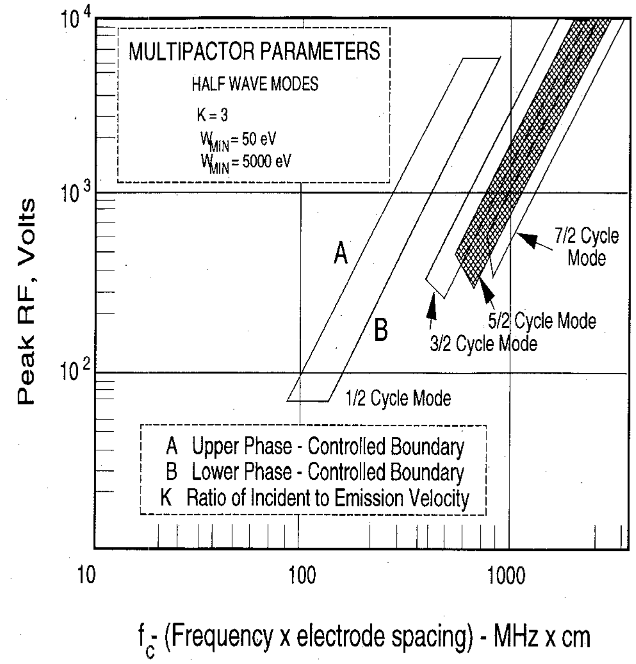


Fig. 13. Possible regions of multipacting between parallel plates.

Dependence on Surface Conditions of Materials

The secondary emission coefficient δ is very sensitive to the relative purity of materials. At low energy levels, there are large variations in δ caused by surface impurities. Most contaminants increase the value of δ and enhance the possibility of multipaction occurrence. Ensuring contamination free implementation and upkeep of high power equipment is critical for space application.

Detection and Prevention of Multipaction

The types of detection methods fall into two categories, local and global. Local detection methods are used close to the point of the actual discharge as in an individual component. For system testing, global detection methods are used which show that a discharge is present somewhere in the high power assembly. The detection methods most often used are the noise floor, forward/reverse power, and balanced phase methods. These methods are described in [49].

In the earlier section on multipaction generation, a number of constraints are described which are essential to initiate and sustain multipaction discharge. Violation of any of those constraints should therefore provide the means to suppress multipaction. Possible ways to prevent the occurrence of this phenomena are enumerated as follows:

Control of frequency-gap product.
Dielectric filling.
Pressurization.
Magnetic or dc bias.
Reduction of surface potential by proper material and processes selection.

A brief description and relative advantages of each of these methods is described in Table IV.

Design Margins in Multipaction

Design margins is a sensitive issue among the customer, prime contractors and suppliers of high power components and subsystems. Owing to the variability of multipaction breakdown, its dependence on workmanship standards, sensitivity to contaminants, complexity of component geometries, change of personnel every few years all contribute to the tendency by the customer to over-specify multipaction margins and test requirements. The prime contractor has often little choice but to accept such specifications for competitive reasons. The supplier of high power equipment is invariably faced with the difficult task of getting such specifications within tight schedules and cost constraints. The optimum choice for all concerned is to adopt certain guidelines in specifying design margins based on theoretical calculations and experimental verifications of a repeatable nature. This aspect was discussed quite extensively at the MTT Workshop [50] and its outcome is summarized here.

The most complex and limiting operating condition with respect to high power is that of multi-carrier operation. This problem can be analyzed by making the following assumptions [48], [49]:

n equally spaced carrier of equal magnitude; and
Over a short duration of time ($\sim$5 ns), the relative phase of the incident carriers at the spacecraft remains unchanged. This condition implies good short-term stability for the local oscillators generating the individual carriers.

Under these assumptions, computed response for the case of 12 equally spaced carriers, 50 MHz apart is shown in

TABLE IV
MULTIPACTION PREVENTION METHODS

Method	Description	Advantages	Disadvantages
Large gap size	The electron transit time is made non-resonant, such that it requires much higher voltage to initiate and sustain multipaction	Simplest mechanical structure	Limitation on RF design, and achievable performance
Gap size below cut-off	The electron transit time is made non-resonant. Electrons are scattered and absorbed, thus preventing the onset of multipaction	No power handling limitations exist due to multipaction breakdown	RF design unsuitable for most practical applications. Achievable values of Q are lower
Dielectric Filling	The mean free path of electrons is reduced by filling the empty spaces of the discharge region with foam or solid dielectric	Simple in theory	Lower the effective Q of the devices. Difficult to apply above 12 GHz
Pressurization	The mean free path of electrons is rendered close to zero by pressurization through an inert gas	Removal of the multipactor mechanism	A possible failure mode if leakage exists. Increase in system weight. System must use hermetically sealed connectors. A source of potential PIM
DC or magnetic bias	A dc bias voltage or applied magnetic field can alter the electron resonance condition to allow higher power handling	Power handling margin improved by ~1 dB, or more. Arching/discharge prevented between the gap	Surface erosion is still possible creating wideband noise and other performance degradation. Performance improvement not significant
Reduced Surface Potential	The secondary emission ratio of the surfaces is spoiled by use of coating of low emission ratio	Simple to apply	Limited applicability due to higher loss

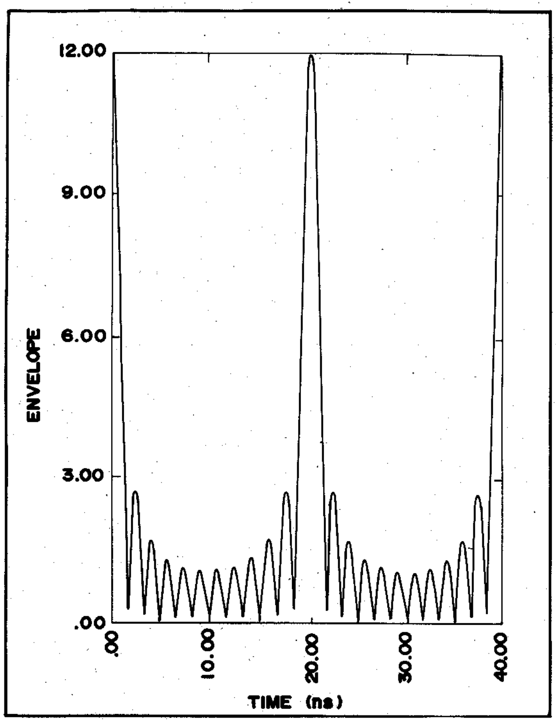


Fig. 14. Computed power envelope of 12 Carriers with 50 MHz spacing.

Fig. 14; the reference for this combined power level is chosen to be when all carriers are in phase. Table V shows the computed power levels for the case of 6, 12, and 100 carriers.

Next, we compare these results with the time it takes to initiate and sustain multipaction. As described elsewhere in this paper, the electrons require a minimum energy level to initiate multipaction. This energy level is built up as the electrons go back and forth between the walls when the resonant condition is satisfied. Thus, a finite transit time envelope is required for electrons to gather enough energy to initiate and sustain multipaction. During the panel discussion at the MTT Workshop [50], an important point was made by Dr. Gerald August of SRI. Based on theoretical considerations and experimental evidence, it is believed that for materials with a secondary coefficient of 2 or less, it takes more than 15 cycles for multipaction to initiate. For first order multipaction, this translates to overall transit times ($\tau = 1/2fo$) of 7.5

TABLE V
COMPUTED POWER LEVELS OF CARRIERS WITH 50 MHz SPACING

Number of Carriers	Dwell Time Envelope (ns)	% of Maximum Power
6	0	100
	0.5	93
	2	0.26
	20	100
12	0	100
	0.5	74
	2	0.025
	20	100
100	0	100
	0.5	0.00
	20	100

Note: The reference for the dwell time of 0 is when all the carriers are in phase. The envelope repeats itself every 20 ns since the carriers are chosen to be 50 MHz apart.

TABLE VI
EXAMPLES OF COM DEV FLIGHT QUALIFIED HIGH POWER EQUIPMENT

Project	Description	Design Approach/ Multipactor Prevention	Power Handing (Peak)
INTELSATs VI and VII	*C*-Band Harmonic Filter	Dielectric Filled Large Gap Separation	> 1.5 kW
INTELSAT VII	*Ku*-Band Transmit-Reject Filter	Large Gap Separation	4 kW
STC DBS	Output Assembly Mux Harmonic Filter	 TE_{114} Filters Stepped Impedance	>2 kW
ERS-1	Output Assembly (5 GHz) Circulator Harmonic Filter Mechanical Switch	 Dielectric Filled Large Gaps Large Gaps	 26 kW Pk 26 kW Pk 45 kW Pk
Mobile Satcom	800 MHz Diplexer	Dielectric Coated Probe Coax-Coupled Design	1 kW
SCS-1	*Ku*-Band Diplexer *Ku*- and *Ka*-Band Isolators	Large Gap Separation Using Ridges and Posts Dielectric Filter	>4 kW >4 kW
Aussat 'B'	*L*-Band Diplexer	Coax-Coupled Design	400 W
Satcom 'K'	12 GHz VPDs	Half-Wave Polarizers	> 500 W
Olympus	18 GHz Diplexer 18 GHz Isolator	Large Gaps Planar Design	180 W

ns at 1 GHz, 1.9 ns at 4 GHz, and 0.6 ns at 12 GHz. These transit time envelopes can now be compared with the 'dwell' times of multi-carrier powers as described in Table V. Based on this analysis, we draw the following conclusions:

(i) For 'n' equally spaced carriers, the 'dwell time' of phased peak power goes down rapidly as the number of carriers increase. Depending upon the actual number of carriers, and the frequency of operation, the 'dwell time' may be less than the time it takes to initiate multipaction. Typical specifications based on additional 3 to 6 dB margin over n^2P_{in} (P_{in} is the power per carrier) are therefore not realistic; and

(ii) For multi-carrier operations when the number of carriers is six or greater, the test levels for multipaction should be specified at power levels no greater than n^2P_{in}. For a smaller number of carriers, the 'dwell time' and frequency of operation should be taken into account when specifying margins for design and testing.

Multipactor Breakdown Levels and Measured Data

Multipaction breakdown has been observed at very low power (of the order of a few watts) levels and more frequently at higher power levels. High Q circuits like filters, multiplexers, transmission lines, switches, and connectors are especially susceptible to this breakdown. This is due to the compact structure of these devices, field enhancement at discontinuities, sharp mismatches outside the band of interest, and the choice of materials.

Table VI describes the high power measured data for a range of microwave equipment [49]. The measured power levels represent the state-of-the-art for high power output circuits required for communications and remote sensing satellites.

Passive Intermodulation (PIM) Consideration for High Power Equipment

It is well known that active devices produce intermodulation (IM) products due to their inherent non-linearity. However, what has not been well understood is that even passive devices like filters and antennas generate lower level IMs, referred to as passive intermodulation products or PIM. In space, PIM was first observed in the late 1960s in the Lincoln Laboratories satellites LES-5 and LES-6. Since then, continuous advances in spacecraft technology requiring a higher degree of channelization and transmitter powers, the nonlinear behavior of passive components has emerged as an important phenomenon, and has become a factor in determining the frequency plan and equipment performance for all high capacity and high power satellite systems.

PIM arises primarily due to the formation of very thin oxide layers on metal surfaces or mechanical imperfections on the joints or both. For metal surfaces separated by 10 to 40 Angstrom thick oxide layers, non-linear electron tunneling through the barrier occurs, giving rise to low level PIM. Nonlinearities also arise due to micro-cracks or voids in metal structures (bulk effects), or due to dirt or metal particles on the surfaces. Build-up of microvoltages at these positions produces micro-currents which also manifest as low level PIMs. The observed values of PIM represent the summation of all these effects. It has been found that for systems requiring 150 to 200 dB isolation between the transmitter and receiver, a nonlinearity of as little a 1 part$/10^{10}$ can be a problem for the system [51].

TABLE VII
AN HISTORICAL PERSPECTIVE OF RF FILTERS AND MUXES FOR *C*-BAND SATCOMS

Type	Program	Typical Mass Per Channel (I/P + O/P) (kg)	Launch Year
Standard Waveguide Invar Filters	*Intelsat IV*	3.6	1971
Thin-Wall Invar Filters	*Anik, Westar*	1.6	1973
Standard GFEC Filters	*First Generation RCA Satcoms*	0.65*	1974–82
Flangeless GFRP Dual-Mode Filters	*Intelsat V*	0.55*	1978
Dual-Mode Super Thin-Wall Invar Filters	*Second Generation RCA Satcoms*	0.83	1982
Axially Mounted Dual-Mode Dielectric Loaded Filters for Input Mux. GFRP Output Mux	*Arabsat*	0.55	1983
Planar Dual-Mode Dielectric Input Mux Invar Output Mux	*Upcoming Satcoms*	0.65	1990s

*Owing to the high cost of manufacturing and elaborate quality control requirements, GFRP filters are rarely employed in the present generation of spacecraft.

The interfaces between high power transmitters and its leakage path to the low noise receiver is where PIM must be controlled to be less than 20 to 30 dB relative to the power of received signals. As a consequence, the high power components in a satellite system must be designed to handle high power and, at the same time, exhibit no or very low PIM. Like multipactor breakdown, PIM is not entirely amenable to theoretical design. It depends upon the material in contact, their surface conditions, workmanship standards and subsequent handling of the equipment. Over the last two decades, industry has built up an extensive experience base in designing, implementing and measuring of high power and low PIM equipment. The most cost effective approach to protect against these twin perils is to force a set of guidelines to be followed at all stages of a satellite program. Such guidelines are summarized in [49]–[52].

XI. CONCLUSION

The past two decades have provided significant advances in reducing the mass and volume of the communications payload, in improving the utilization of the frequency spectrum and enhancing transponder performance in terms of more efficient delivery of higher EIRPs. The net effect has been the lowering of the cost of a satellite channel.

A significant portion of these improvements has come via the innovations in the microwave filters and multiplexer technology. The dramatic reduction in mass since the launch of first commercial communications satellites to the present generation is shown in Table VII. Looking ahead, it is quite likely that output muxes using quarter or half-cut dielectric image resonators would be qualified for *C*-band satellite systems. This would provide additional mass savings.

For *Ku*-band satellite systems, input mux using dielectric resonators have been qualified. However, the average unloaded Q achieved is significantly less than the dual-mode TE_{113} filters. Improvements in dielectric materials and assembly techniques would likely increase the value of unloaded Q to an acceptable level. Under such an advance, significant savings in mass (3 to 4 kg) and volume would be achieved. Owing to the small size of dielectric resonators coupled with much higher power requirements, it is unlikely that output muxes at *Ku*-band would use dielectric technology.

In the distant future, superconductivity offers an attractive option for realizing microwave filters and multiplexing networks for space application. Such a technology would enable construction of planar thin film microstrip type filters having a performance equivalent to that which can be achieved by present technology. That would provide a large reduction in mass and volume of the muxes. This is an area of intense R&D worldwide. It would be an exciting event should this technology come to fruition for spacecraft multiplexers. Another technique that has received sporadic attention for realizing muxes is the surface acoustic wave (SAW) technology. This technology can provide extremely sharp filters and inherent linear phase in the passband in a very small size and mass. It has the disadvantage that its preferred band of operation is from 50 to 500 MHz and the passband has large group delay ripples ($\cong$ 25 ns) of short frequency intervals. It is generally believed that such fast varying ripples do not cause much distortion for most types of traffic and thus would be acceptable. The use of the preferred frequency band for such devices would require double conversion for *C*- or *Ku*-band satellite systems and probably a single conversion for the mobile *L*-band satellite systems. SAW based IF processors including banks of SAW channel filters are being designed and qualified for the next generation of Inmarsat satellites [53]. This technology has good potential for use in the next generation of mobile satellites, especially for narrow band applications.

Acknowledgment

The authors acknowledge the contributions of many of their colleagues over the past two decades in advancing the filter and multiplexing technology for space applications. Special thanks is owed to Mr. Val O'Donovan, Chairman of the Board and CEO of COM DEV whose leadership has been instrumental in the provision of a stimulating environment in the company that led to many innovations.

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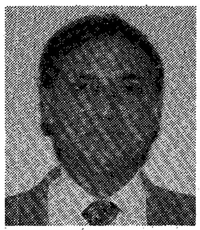

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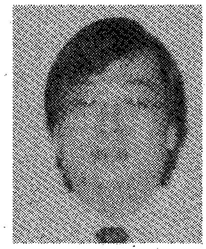
Richard Cameron, photograph and biography not available at the time of publication.



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